(handwritten) 11 ADDRESS × 44   45 50 IC
14 1/6 30   54 CODE CONVERTER

Elektor Electronics

Dual-tone multi-frequency (DTMF) system decoder

disable / enable   connected   DTMF data

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May 1989
Volume 15
Number 167

In next month's issue:

- In-circuit transistor tester
- In-line RS232 monitor
- Guitar amplifier
- CMOS switches for audio applications
- MIDI keyboard
- Diodes and their uses
- Disconic antenna
- Echo Unit
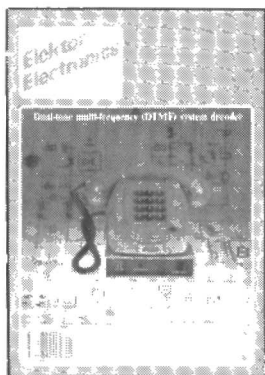
Front cover

The Dual-Tone Multi-frequency Decoder shown here underneath a standard telephone set is a state-of-the-art design based on Teltone Corporation's single-chip receiver Type M957.
The DTMF dialling scheme for telephone networks was developed by Bell Laboratories and introduced in the United States in the mid-1960s as an alternative to pulse dialling. Offering increased speed, improved reliablility, and the convenience of end-to-end signalling, DTMF has since been adopted as standard and recommended for use by telecommunication organizations such as CCITT, CEPT, NTTPC and many others around the world.

The digital model train
p. 16

Analogue multimeter
p. 18

Radio Data System Demodulator
p. 32

A closer look at the transputer
p. 39

Britain in the RACE

The award of almost £80 million worth of orders to British firms in the final phase of the European Research into Advanced Communications Equipment programme is encouraging news. It means that British companies have gained more than a quarter of the last allocation which covers some 40 different projects. The total budget for the research programme amounts to almost £900 million.

Since the programme is a cost-shared one, the latest award means that the 40-odd British companies involved have to invest a combined sum of close to £80 million in the various research projects. Virtually all the important telecommunications companies in the United Kingdom, such as STC, GEC, Plessey, British Telecom, Philips, BICC Cables, and a host of others are among those to receive a share of the funding.

A recent report on the likely effect of the programme on the British telecommunications industry says that full exploitation of the results of the research projects would require the participating companies to invest an amount exceeding £300 billion during the last ten years of this century.

<div align="center">*   *   *   *   *   *   *</div>

Sound future for SMT

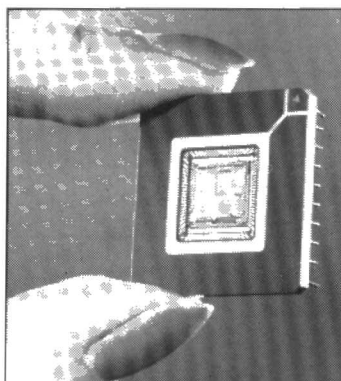Although there are still some who doubt the viability of Surface Mount Technology, there is ample evidence that the use of surface mount components is growing rapidly throughout the industrialized world,

None the less, there remain a number of problems of which the most serious is probably the absence of agreed international standards of assembly and inspection. Another is the difficulty of visual inspection (automated inspection systems can not — yet — take over completely from the human inspector), which stretches human capabilities to their limit (think, for instance, of the thousands of solder joints on a single Eurocard).

However, the first step to the solution of a problem is recognition of the problem and it is widely accepted that most pitfalls associated with surface mount technology have been recognized. In any case, the worldwide growth of SMT speaks for itself. If it were not a viable production method offering many advantages, it would have died a natural death by now.

<div align="center">*   *   *   *   *   *   *</div>

Change of address

From 3 May 1989 our editorial address will be

<div align="center">

ELEKTOR ELECTRONICS (Publishing)
Down House • Broomhill Road • LONDON SW18 4JQ
Telephone 01-877 1688 • Telex 917003 • Fax 01-874 9153

</div>

Our European (production) Offices remain at

<div align="center">

P.O. Box 75 • 6190 AB Beek (L) • The Netherlands
Telephone +31 4490 89444 • Telex 56617 • Fax +31 4490 70161

</div>

ELECTRONICS SCENE

## Satellite broadcasting

The current and future state of satellite broadcasting will be the focus for discussion at the sixth annual European Satellite Broadcasting Conference to be held on 22–23 June at the Queen Elizabeth II Conference Centre in London. It will bring together the leading problem solvers and strategic planners in this volatile and rapidly expanding business.

Further information on the event may be obtained from the organizers: Blenheim Online • Ashill Drive • PINNER HA5 2AE • Telephone 01-868 4466.

Ethernet transceiver

The Ethernet Transceiver from Ringlan Components is approved to IEEE 802.3
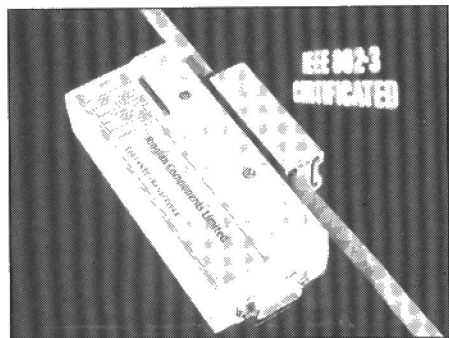
The unit incorporates a NETSCAN facility that allows fault diagnosis without disruption to users on the network.

Full details from Ringlan Components •58 Burkitt Road • Earlstrees Industrial Estate • Corby NN17 2DT • Tel. (0536) 400517.

Airline videos

The world's first in-flight personalized video systems are being evaluated by British Airways on board its Boeing 747 aircraft. The systems allow passengers the freedom to decide what they want to watch, when they want to watch, and gives them the control to start, stop, rewind and view again. The aircraft will have a library of about 50 video cassettes on board, ranging from comedy to drama, classics or documentary.

New test centre for System X

Telecommunications administrations throughout the world will benefit from a new 1,858 m² test facility being established by GEC Plessey Telecommunications The centre, which is due to come into operation later this year, will house ten feasibility model System X digital telephone exchanges, making it the largest of its kind in the United Kingdom and one of the largest in the world.

Automatic announcers

An announcing system in which solid-state electronics are used to reproduce 'real' voice has been introduced by P A Communications.

The 'Parva' announcers may be used as a single-line telephone answering system, for multiple-line answering, or for announcements in railway stations. Each announcer can be linked to up to 100 telephone lines.

Full details from P A Communications Ltd • 10 Darin Court • Crownhill • Milton Keynes MK8 0AD • Tel. (0908) 567248.

Call-alert adaptor for cellular phones

Drivers with cellular telephones can be alerted to incoming calls while away from their vehicles by an adaptor that switches on a warning device.

The 'Call-alert Sounder' from ACC-CELL, is driven by pulses from a current-sensing adaptor actuated by an incoming call. It is intended for use with cellular telephones that do not have an in-built call-alert facility.

The unit is produced by ACC-CELL Ltd • Apex House • Kingsfield Lane • Longwell Green • Bristol BS15 6FL • Telephone (0272) 352261.
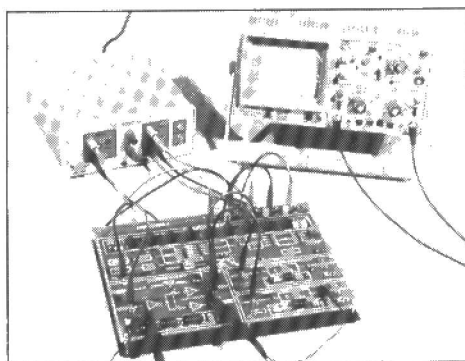
**The CT1 module in the 'Modicom' range of training boards from LJ Electronics is for hands-on instruction of student electronics engineers and technicians in digital communications. There are three units in the range, each of which is available separately with individual user manuals or as a package with complete study curriculum.
LJ Electronics • Francis Way • Bowthorpe Industrial estate •Norwich NR5 9JA.**

Mini aerial for satellite reception

Mr Patrick May, a senior lecturer in the engineering science department at Exeter University has perfected a new small aerial and simple frequency changer that promises to replace many of the large and unsightly dish aerials used to receive satellite TV programmes.

The aluminium tube antenna can be either a 'V' shape, typically 609 mm tall, or rhombic. It is of the 'travelling wave' type capable of receiving transmissions over a wide range.

Video standards translator

A translator that makes the transfer of video signals between VHS and Umatic recording machines possible has been developed by Midlands Video Systems. The TT-880-P converts the commands from a Sony RM-440 Umatic editing controller into commands recognized by a Panasonic industrial VHS recorder. Signals fed back from a Panasonic machine are translated into Sony commands.

Midlands Video Systems Ltd • 3a Attenborough Lane • Chilwell • Nottingham NG9 5JN.

Secure voice communications

The 'BA 1772' narrow-band secure voice unit from MEL offers secure voice communication on HF, VHF and UHF radio networks. It consists of a voice coder (vocoder), an encryption unit and a modem.
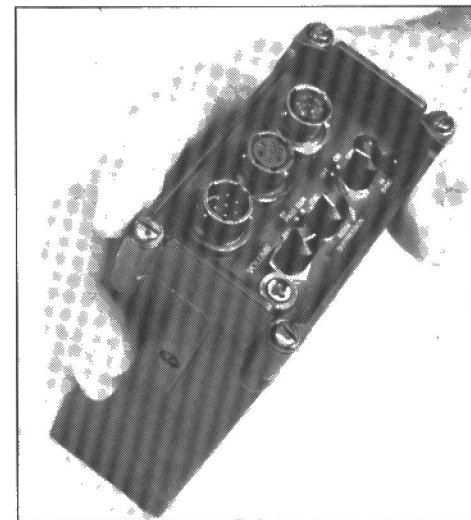
Suitable for use with backpack, mobile or fixed radio stations, it operates on a conventional voice channel with no modification of the transmitter-receiver, and will operate over all types of transmission path. MEL • Manor Royal • Crawley RH10 2PZ Telephone (0293) 28787.

Fax sent direct by computers

An electronic interface from Hasler allows text and diagrams to be translated into the signals sent by a fax machine without the intervention of a scanning device. It allows text images sent by a computer to be received by any Group 3 facsimile receiver or by another computer similarly equipped.

The 'Hasler Fax Unit' is said to offer a much clearer image than that from a scanning fax transmitter since the image is translated direct by the interface.

Hasler (Great Britain) Ltd • Commerce Way • Croydon CR0 4XA • Telephone 01-680 6050.

Computerized film and TV making

A complete range of computerized motion-control equipment together with a custom design and manufacturing service is available from Mark Roberts Film Services. The company's range of products includes computer systems, model rigs, special-purpose rostrums, fade/dissolve control units, automatic focus units, camera/projector motors and custom designs. The company can also update and computerize existing systems or it can design, supply and manufacture complete systems to customer specifications.

Mark Roberts Film Services • Birches Industrial Estate • Imberhorne Lane • East Grinstead RH19 1XZ • Tel. (0342) 313522

Digital speech recorder and analyser

A digital recording system that makes the transcription of recorded speech easier, quicker and more accurate has been produced by Racal Recorders. The system makes it possible to extract the intelligible speech from faint or crackly distant radio or telecommunications transmissions. The voice information is recorded digitally on to hard memory disks. With the aid of advanced software any point on the disk may be indexed precisely so that it can be returned to and repeated as often as required. The use of digital recording ensures that there is no loss of fidelity throughout the record / replay process.

Racal Recorders Ltd • Hardley Industrial Estate • Hythe • Southampton SO4 6ZH • Telephone (0703) 843265.

Cable-end system

A multicore-cable termination that enables users to assemble cable ends on their own premises with the minimum of tooling is available from Icore. 'Optimum 11' solves the problem of joining multicore cable to multipin connectors and is easy to install. It replaces the last few inches of multicore cable with a flexible conduit. This avoids the installation and vibration problems that occur when the normally stiff cable is transferred into the back of a multipin connector.

Icore International Ltd • Leigh Road •Slough SL1 4BB • Tel. (0753) 74134.

Compact telex tester

The 'T 300 Telex Analyser', a new test equipment for the engineer in the field, from Componedex is compatible with SPC and Strowger systems. It is able to decode Baudot, Hex, and ASCII protocols and is designed to cope with the higher speeds of the next generation of telex machines – up to 300 baud. It incorporates in a single unit functions that have hitherto required separate items of test equipment, and it will test power levels, establish contact with a remote terminal and discover its identity (handshaking), and locate breaks in the line between base and exchange.

Componedex Ltd • 21 Alston Drive • Bradwell Abbey • Milton Keynes MK13 9HA • Telephone (0908) 322177.

ELECTRONICS SCENE

New method of 'turning around' live radio material

BBC engineers have devised a new method of using digital technology to record, store and replay incoming live material with just one machine. This development paves the way for a more efficient means of 'turning around' live radio material

Wideband fibre optic link

The first ultra-wideband fibre-optic link in Europe has been produced by Marconi. The 2–20 GHz, 1 km long transmission line represents a new technology offering vastly superior bandwidth and insertion loss compared with existing coaxial cable or waveguide, and also gives immunity to electrical interference.

Bandwidth capability is equivalent to 3000 television channels or six million telephone conversations transmitted simultaneously down a single strand of optical fibre. In addition, many optical carriers can be multiplexed down the same fibre on slightly different wavelengths.

FIBER OPTIC COMMUNICATIONS COMPONENTS MARKET IN EUROPE - 1993

SOURCE: Frost and Sullivan, Ltd.
Report #E1012

Cabled Fiber $476M
Test Equipment $123M
Splicers/Kits $28M
TDM Units $33M
Couplers/WDM $28M
Connectors $56M
Tx/Rx/Modems $167M

0 50 100 150 200 250 300 350 400 450 500
$Million
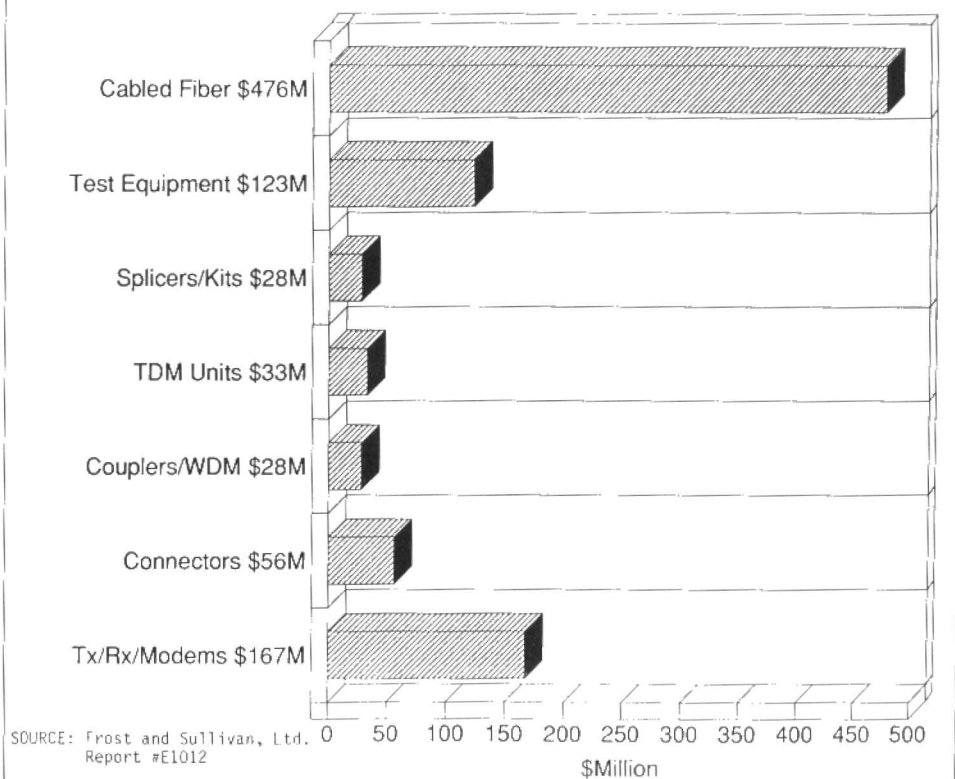
According to a report from Frost & Sullivan, there is enough fibre optic cable in Europe to circle the earth twice – 90,000 km. A new study by the company forecasts that sales of fibre optic components will double between 1988 and 1993.

INTERMEDIATE PROJECT

A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary understanding and knowledge of electronic engineering. Each project in the series will be based on inexpensive and commonly available components.

3b. Sine wave convertor

by K. Walters

The simple triangular-to-sine wave converter described here complements the basic signal source described last month.

A function generator is not complete without an output that supplies a stable, sinusoidal, waveform of adjustable frequency and amplitude. Fortunately, a sinusoidal signal is fairly simple to generate from a triangular signal, which is supplied by the basic function generator module described in last month's instalment of the *Intermediate Project* series. As already mentioned there, the function generator is a user-configurable design, to which we are now about to add a further module: a triangular-to-sine wave convertor.

presence of a silicon diode in the circuit. One of the basic features of a diode is that it starts to conduct at a forward voltage, U_d, of about 0.6 V, and passes more current, I_d, as the voltage increases (see the U-V curve in Fig. 2). A reasonable approximation of a sine wave would, therefore, be achieved by applying a triangular signal to a circuit as shown in Fig. 3.

The actual convertor is formed by R_1 and anti-parallel diodes D_1-D_2. Opamp IC_1 functions as a buffer to prevent the sine wave being distorted by the load imped-

tent the shape of the output voltage.

The above operation is ideal, and only achieved in the theoretical case where the p.d. is *not loaded*. In all other cases, the load draws current from the p.d., so that the current through R_1 becomes

Fig. 2. Basic voltage-current (V-I) characteristic of a silicon diode. Note that I_d rises sharply above 0.6 V.

Fig. 1. The sine wave converter translates a triangular input voltage (left-hand photograph) into a sinousoidal waveform (right-hand photograph).

First attempt: the diode limiter

The differences between a triangular and a sinusoidal signal are apparent from the oscilloscope photographs in Fig. 1 (the vertical sensitivity of the scope is set to 1 V per division). The signals shown are generated by the present convertor.

Looking at the curves, it would appear that the sine wave is simply a triangular signal with rounded, rather than sharp, inflection points. There is virtually no difference between the signals around the zero-crossing. The curve of the sine wave, however, becomes rounded when the instantaneous amplitude rises above 0.8 V or so. This can not but point to the

ance, as will be seen later.

Series resistor R_1 and diodes D_1-D_2 form a special type of potential divider (p.d.) for the positive and negative excursion of the input voltage. The output voltage of the p.d. is a function of the current passed by the diodes. This current in turn depends on three factors: the instantaneous input voltage, the value of R_1, and the U-I characteristic of the diodes used. Since the input voltage rises and falls linearly, and R_1 is constant, the U-I characteristic is the main factor to determine the current that flows as a result of the instantaneous input voltage. Hence, the characteristic determines to a large ex-

Fig. 3. The basis of a triangular-to-sine wave converter is formed by a voltage-dependent potential divider and a buffer.

greater than the forward current through either diode. The additional current changes linearly with the input voltage, giving rise to a triangular voltage on R_1. Obviously, this voltage causes severe distortion of the sine wave.

The opamp connected to the p.d. has an input current of almost nought, so that it supplies an output voltage that is virtually identical to the ideal voltage obtained from the p.d. The opamp so obviates load effects on the p.d. and at the same time provides sufficient drive power.

Feedback does it

The above discussion could lead to the assumption that the circuit of Fig. 3 is the perfect sine wave convertor. Unfortunately, this is not the case.
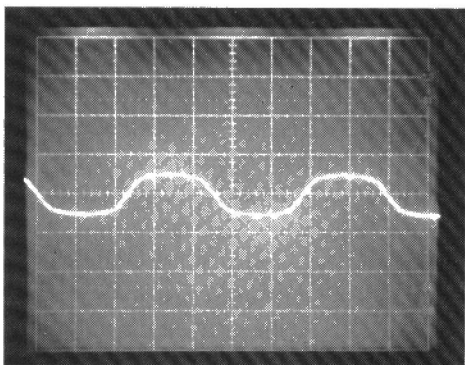
Fig. 4. The waveform obtained from the basic circuit of Fig. 3 can hardly be called a sine wave.

The scope display in Fig. 4 shows the waveform supplied by the circuit of Fig. 3. The signal has flattened tops and resembles only vaguely a sine wave. Apparently, the voltage is limited too severely by the diodes.

A better sine wave is obtained by extending the circuit of Fig. 3 with a second potential divider that provides feedback of a portion of the output voltage. The practical circuit is shown in Fig. 5: the feedback arrangement 'raises' the ground reference of the diodes, and effectively reduces the voltage applied across them to a level that results in less severe limiting.

The operation of this circuit is basically as follows: an increasing output voltage gives rise to a voltage between the wiper of P_1 and ground. The voltage between R_1 and ground will, however, not change much with respect to the situation in Fig. 3, so that a smaller voltage is applied across the diodes. These are, as a result, operated over a 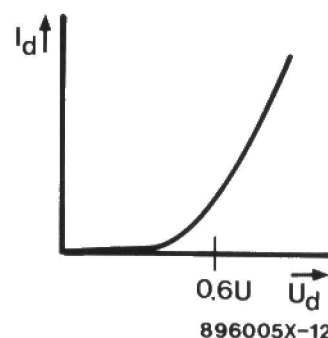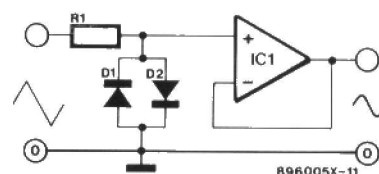relatively small part of their U-I characteristic. With P_1 adjusted correctly, the diode voltage will not rise very far above the conduct level around 0.6 V. The use of feedback results in a sine wave with acceptable distortion, as already shown in Fig. 1.

A low-pass filter, R_3-C_3, is provided at
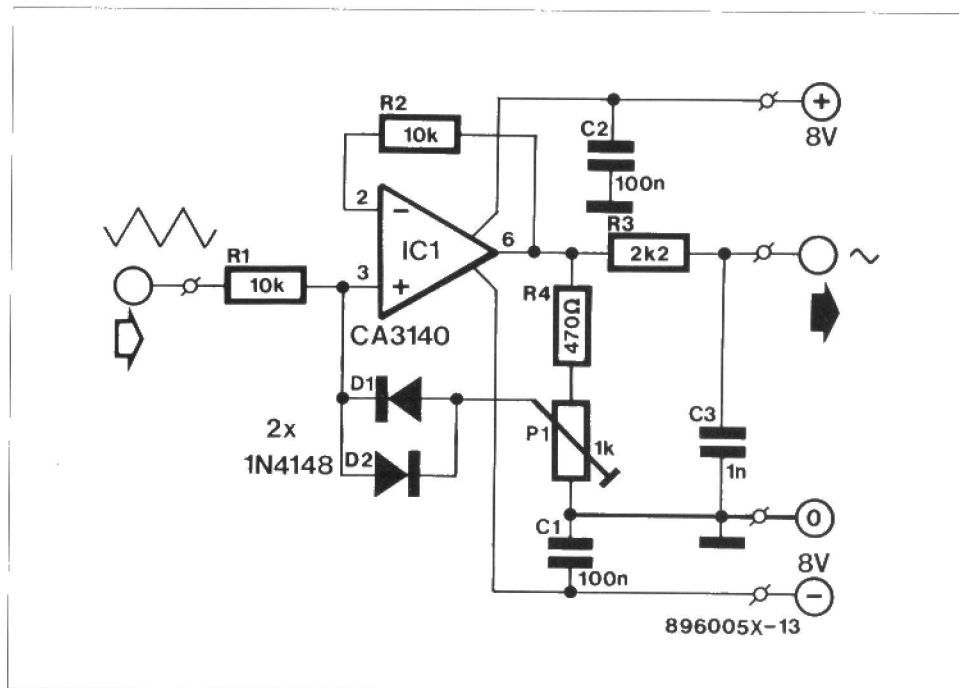
Fig. 5. Circuit diagram of the triangular-to-sine wave convertor, the second module in the low-budget function generator.

the output of the triangular-to-sine wave convertor. This filter suppresses harmonics and helps to improve the shape of the sine wave. Feedback resistor R_2 is included to stabilize the opamp.

Building and testing

The component mounting plan of Fig. 6 shows that the sine wave convertor occupies only half the available surface on Universal Prototyping Board Size-1 (UPBS-1) used for most circuits discussed in this series of articles.

Construction should not present prob-

lems because relatively few components and wire links are used.

The circuit is connected to a symmetrical 8 V supply (or two 9 V PP3 batteries). Set P_1 to the centre of its travel. Connect the input to the triangular wave output of basic generator module described last month. Connect the output of the convertor to the input of an oscilloscope. Watch the scope display while carefully adjusting P_1 until a reasonably shaped sine wave is obtained. If a dual-beam scope is available, the sine wave may be compared with the secondary voltage from a mains transformer.

The last two modules in the function generator, the output amplifier and the power supply, will be described in a forthcoming instalment of the *Intermediate Project* series.
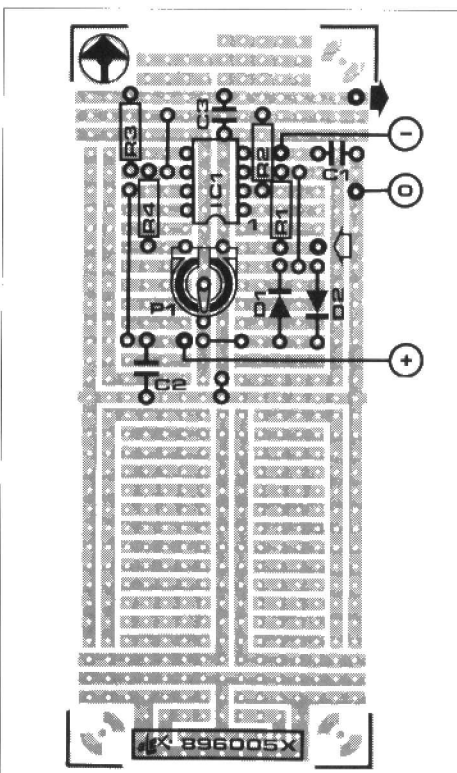
Fig. 6. Construction on board UPBS-1.

Parts list

Resistors (±5%):
R_1;R_2=10K
R_3=2K2
R_4=470R
P_1=1K0 preset for horizontal mounting

Capacitors:
C_1;C_2=100n
C_3=1n0

Semiconductors:
D_1;D_2=1N4148
IC_1=CA3130

Miscellaneous:
8-pin IC socket.
PCB Type UPBS-1 (see Readers Services page).

THE DIGITAL MODEL TRAIN – PART 4

by T. Wigmore

Universal signals and switching decoder
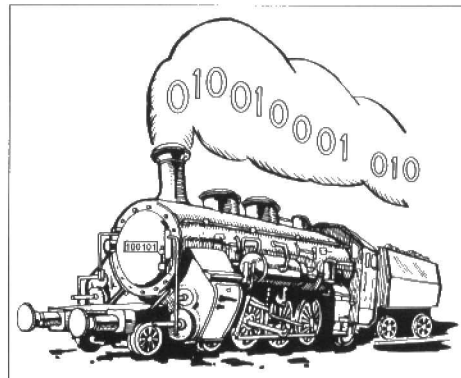
The points (turnout) decoder described in Part 1 of this series (Ref. 1) is a circuit that provides short pulses to its output that may be used for the control of turnouts (points) and other electromagnetic elements of the model railway. That decoder does not provide a momentary but a continuous contact. It is really a remotely controlled four-way toggle switch with change-over contacts.

Although the decoder described in this instalment enables all kinds of element to be switched, the design is aimed primarily at (non-electromagnetic) light signals. It provides separate switched outputs to which LEDs or conventional light bulbs may be connected. If desired, it is possible to mount relays on the PCB. These would, for instance, enable switching the track voltage at the same time as the signals, in effect providing the basis for hardware-based block protection.

Four sets of light signals may be connected to a decoder. If relays are used, it is possible to switch circuits that are of necessity electrically isolated from the track.

The present decoder uses the same address location as the points decoder. It is suitable for use with the Märklin HO system and the "Elektor Electronics Digital Train System – EDiTS" to be published in future instalments in this series, commencing next month.

Circuit description

The circuit diagram of the decoder, see Fig. 28, shows some similarity with that of the points decoder in Fig. 4. Use is made again of the MC145027 to analyse the serial data emanating from the rails into adresses and control data. The trinary address, A1–A4, is set with the aid of jumpers or wire links. A5 is an address bit, but since it is permanently connected to ground it is always logic 0. None the less, provision has been made to give this bit another value in the future to make it possible, if required, to give the present decoder a different address location. The four data bits

ara available at outputs D6–D9.

Three-to-eight decoder IC2 generates a set or reset signal for one of the R–S bistables. To obviate critical races, a set or reset pulse is passed on only after the Valid-Transmission signal has become active. The last switching position of the four channels is stored by IC3, a quadruple R–S bistable.

The output stages are based on Type ULN 2004 darlington arrays. The signal lights may be connected direct to the darlington outputs, which are switched with respect to the positive supply line (about 20 V). For convenience, series resistors (R7–R10) have already been provided in the common lines, so that signals using LEDs may be connected without further work. If conventional light bulbs or other elements are to be connected, the value of the resistors may be adapted as appropriate or they may be replaced by wire links to ensure that the connected element is provided with the correct voltage.
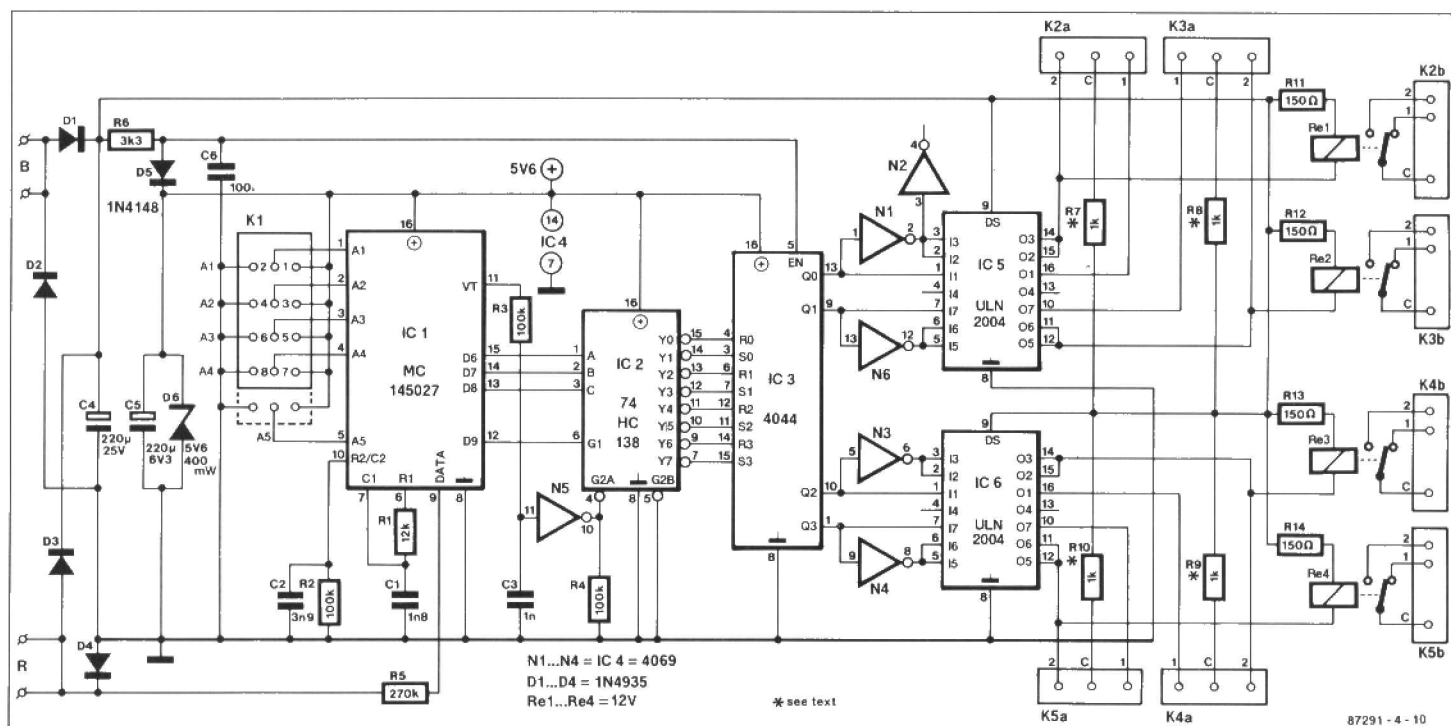
Fig. 28. Circuit diagram of the universal signals and switching decoder.

Parts list

Resistors:
R1 = 12 k
R2–R4 = 100 k
R5 = 270 k
R6 = 3k3
R7–R9 = 1 k*
R11–R14 = 150 R

Capacitors:
C1 = 1n8
C2 = 3n9
C3 = 1 n
C4 = 220 μ/25 V radial type
C5 = 220 μ/6V3 radial type
C6 = 100 n

Semiconductors:
D1–D4 = 1N4935**
D5 = 1N4148
D6 = 5V6/400 mW
IC1 = MC145027
IC2 = 74HC138
IC3 = 4044
IC4 = 4069
IC5; IC6 = ULN2004

Miscellaneous:
Re1–Re4 = 12 V relay for PCB mounting
       with single change-over contact

K1 = 3×4 terminal block for PCB
       mounting with 4 shorting plugs
PCB 884012

*If light bulbs are used instead of LEDs,
the value of these resistors must be
adapted as appropriate.

**Other types in the 1N493X series may
also be used
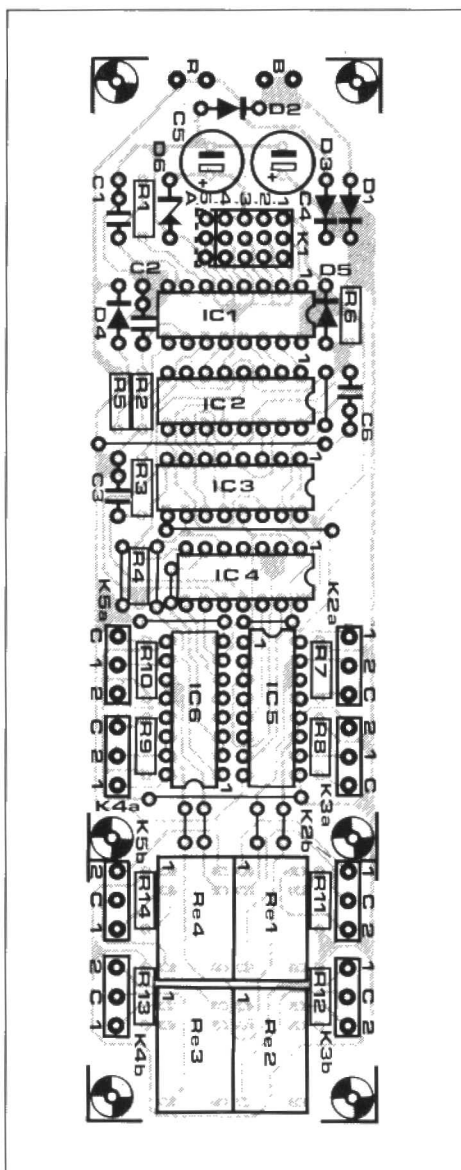
Fig. 29. The printed-circuit board for the universal signals and switching decoder.

Fig. 30. Illustrating the manner in which signals with three lights may be connected

87291 - 4 - 12

Fig. 31. Numbering of the jumpers or wire links on connector K1.

The maximum current drawn by IC5 and IC6 is 1 A (actually limited by the rectifier). Each separate channel may switch not more than 500 mA. If higher currents are required, or if the particular circuit is to be electrically isolated from the rails, relays must be used. These relays may also serve to provide automatic control of trains if the present decoder is used for the operation of signals.

The usual freewheeling diodes for the relay coils are contained in the darlington arrays. Resistors R11–R14 serve to limit the current drawn by the relay coils. This is necessary because the supply voltage (18–20 V) is a little high for 12 V relays.

If the supply is interrupted (for instance because of a short-circuit between the rails), the last data will be retained for 20–30 seconds thanks to C5. As long as the supply is restored fairly quickly, operation of signals and other connected elements will be recommenced from the operational condition that existed immediately prior to the supply interruption.

Taking the decoder into use

Terminals B and R are connected to the rail that is supplied by the brown and red wires respectively of the digital system (Märklin colour coding).

Low-current (<500 mA) elements, such as signals, may be connected direct to the 'a' (darlington) outputs. The total current drawn by such elements must not exceed 1 A at any one time. If LEDs are used, series resistors R7–R10 must be fitted.

When the red button on the keyboard is pressed, outputs marked '2' are active and the relevant relay is energized.

Take care that signals controlled by the present decoder are isolated from the rails. Signals with three lights should be connected in accordance with Fig. 30.
Elements, whether electrically isolated from the rails or not, that draw a current exceeding 500 mA must be connected to the 'b' (relay) outputs. The switching current should not exceed 1 A. These outputs may also be used for block control.

Address settings are as given in Table 1 (Ref. 1). A closed DIL switch in that table corresponds to a jumper or wire link in the present decoder. The numbering of these links is shown in Fig. 31. No jumper or link needs to be placed for address 5, because a wire link to ground has already been provided on the PCB.
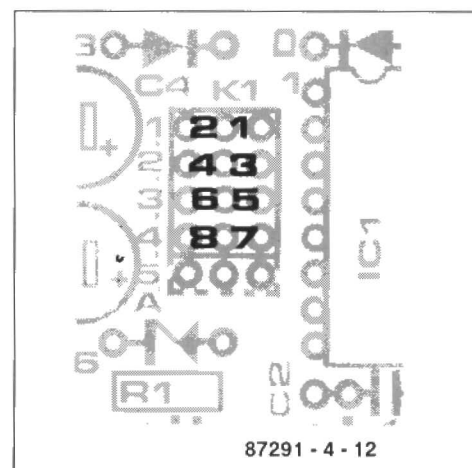
References:
Elektor Electronics, February 1989, p. 42

Next month we commence with the Elektor Electronics Digital Train System.

ANALOGUE MULTIMETER

by T. Giffard

An advanced electronic multimeter with no fewer than 80
ranges and easily reproducible accuracy.

An important consideration with home construction of test equipment is invariably the achievable accuracy and the dependence thereof on the calibration procedure. The analogue multimeter described here has no alignment points simply because its accuracy is derived mainly from close-tolerance (0.1%) resistors. These components have long been outside the reach of the hobbyist, but are now gradually becoming available in small quantities and at reasonable cost. The 0.1% resistors used in the present meter may be replaced with less expensive 1% types at the cost of some accuracy.

One feature that makes the analogue multimeter a particularly interesting test instrument is the large number of ranges available for measuring voltage, current and resistance: there are no fewer than eighty.

Analogue *versus* digital

The choice between a digital and an analogue multimeter is often made on the basis of personal preference. Although digital multimeters are now commonly found in almost any electronics workshop, the analogue meter holds its own thanks to its ability to show the *trend* of a measured quantity. This feature is especially useful for alignment purposes, since it is easier to observe the movement of a needle than to resolve four or more rolling digits and, possibly, a shifting decimal point on a digital read-out. Another benefit of the analogue meter scale is that it allows a simple way of indicating tolerance and high/low ranges in coloured areas on the meter scale.

Probably because of the 'absolute' display indication, digital multimeters are often, but in most cases wrongly, credited with better accuracy than analogue meters. Many electronics enthusiasts are convinced that any type of display, whether LED or LCD-based, rules out read-out errors because the indication itself is not liable to misinterpretation. They forget, however, that the final accuracy of both the digital and the analogue multimeter is determined by the electronics inside, not the read-out only. In fact, the additional error introduced by carefully reading the measured value on a good analogue instrument with a class-1.5 or better mirror-type moving-coil meter is nearly always smaller than that introduced by any component with 1% tolerance. As examples, 3 V is easily resolved on a good analogue scale of 300 V, and 0.1 V on a scale of 10 V.

Combined R-V-I measurement

The design of the analogue multimeter is based on a voltage meter with a basic full-scale deflection (f.s.d.) sensitivity of 1 mV. There are several ways of making such a meter circuit: either a separate voltage amplifier could be added for the 1 mV range, or the entire instrument could be designed such that all ranges are scaled down to the most sensitive range of 1 mV.

The multimeter discussed here is based on the latter principle. The main reason to opt for a common range of 1 mV is, perhaps surprisingly, a very practical one: it is the use of a 12-way rotary range switch with one pole. From a point of view of design, a double-pole 12-way switch might have been a better choice. Such a switch is not used here, however, because of its higher price and the requirement for additional components.

Voltage measurement
The voltage ranges are created conven-

tionally with the aid of a switchable voltage divider as shown in Fig. 1a. The voltage divider is designed such that a scale factor of 10 dB (approx. 3 times or √10) is obtained. The analogue multimeter has 12 voltage ranges from 1 mV f.s.d. to 300 V f.s.d. The fixed voltage divider ensures that the meter has a constant input resistance on all ranges. This means that range switching does not affect the measurement as it does on many non-electronic analogue multimeters (with *non-electronic* we mean: not using active devices such as FETs, bipolar transistors or opamps).
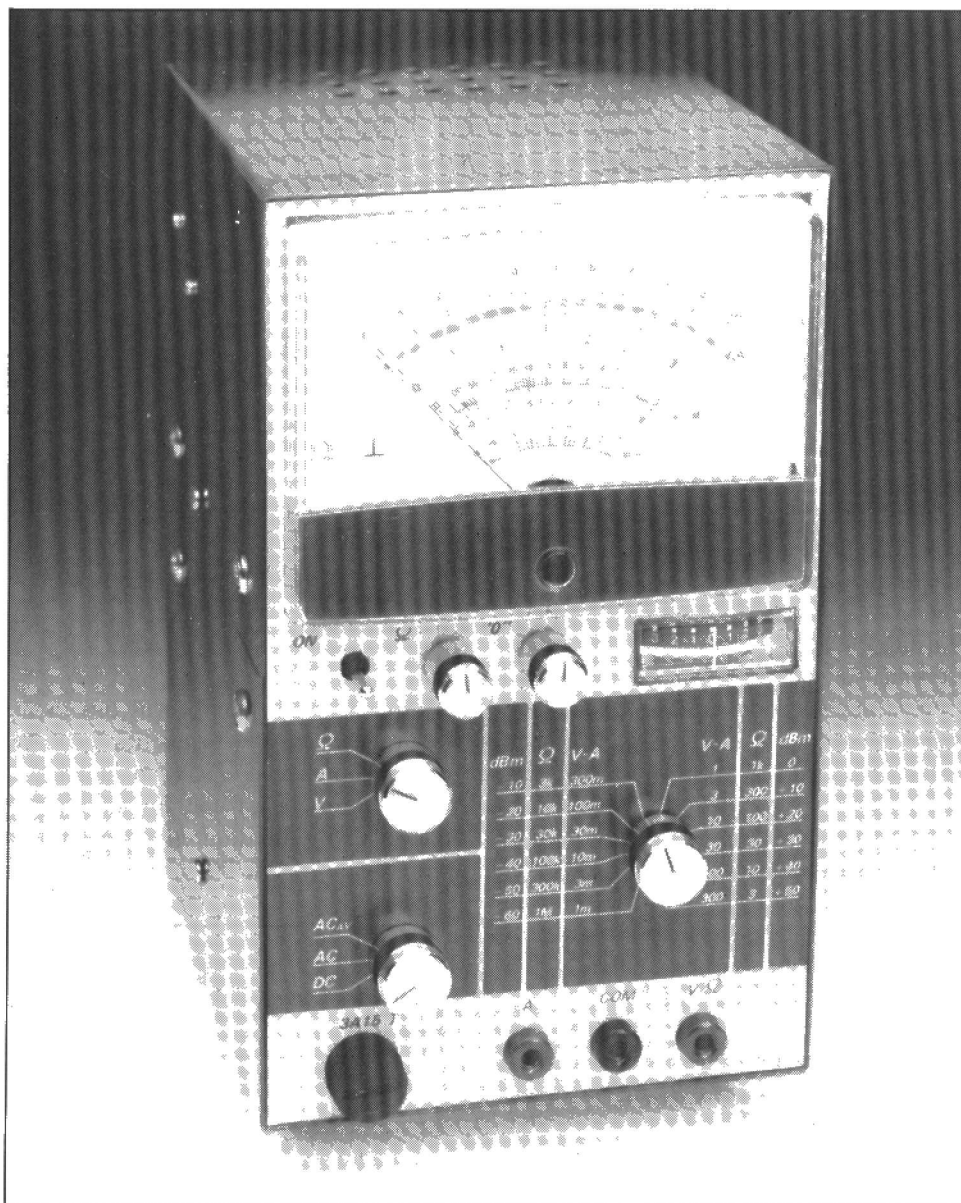
Current measurement

A shunt resistor translates measured current flow into a proportional voltage as shown in Fig. 1b. The shunt resistor has the ideal value of 1 Ω so that the measured voltage can be taken to represent the measured current without the need of a correcting factor.

Resistance measurement

Resistance can be measured by incorporating the unknown resistor, R_x, in a voltage divider as shown in Fig. 1c. The second resistor is an adjustable reference type that forms part of the voltage divider. The voltage across this resistor is proportional to the value of the measured resistor. Resistance measurement with this method has the disadvantage of requiring a non-linear scale, but is still attractive because it is simple to combine with the V (voltage) and I (current) ranges of the multimeter.

The voltage divider is supplied with a bias voltage of 10 mV. This limits the voltage measured across the reference resistor to a maximum of 10 mV, and requires the sensitivity of the basic millivolt meter to be lowered from 1 mV to 10 mV for resistance measurements.

Seven sub-circuits: one multimeter

Combining the three previously mentioned measurement principles, showing the internal configuration of the 1 mV meter amplifier, and adding a power supply to the whole results in the circuit diagram of Fig. 2.

Rotary switch S_2 selects between R, V and I measurement.

The principles illustrated in Fig. 1 are fairly simple to recover from the circuit diagram. The adjustable voltage divider of Figs. 1a and 1b, and the adjustable resistor of Fig. 1c, are formed by rotary switch S_1 and resistors R_1 through R_{14}. The resistors serve the double function of voltage divider and references for the Ω ranges. This arrangement not only saves resistors, but also allows a single-pole rotary switch to be used. R_1 through R_{14} and all other 0.1% resistors are standard values from the E96-series. In most ranges, the final accuracy of the

Analogue Multimeter

SPECIFICATION

Voltage measurement:
Direct voltage
Alternating voltage: effective value of sinusoidal
  waveforms (AC$_{rms}$)
Alternating voltage: average value (AC$_{AV}$)

Input impedance: 1 MΩ
Frequency range: DC...20 kHz
Protection: max. 300 V$_{rms}$

Ranges:
1 mV; 3 mV; 10 mV; 30 mV; 100 mV; 300 mV; 1 V; 3 V; 10 V; 30 V; 100 V; 300 V

Current measurement:
Direct current (DC)
Alternating current: effective value of sinusoidal
  waveforms (AC$_{rms}$)
Alternating current: average value (AC$_{AV}$)

Internal resistance: 1 Ω
Frequency range: DC...20 kHz
Protection: fuse 3.15 A slow

Ranges:
1 mA; 3 mA; 10 mA; 30 mA; 100 mA; 300 mA; 1 A; 3 A

Resistance measurement:
Test voltage: 10 mV
Protection: none

Ranges:
3 Ω; 10 Ω; 30 Ω; 100 Ω; 300 Ω; 1 kΩ; 3 kΩ; 10 kΩ; 30 kΩ; 100 kΩ; 300 kΩ; 1 MΩ

Signal level measurement:
−60 to + 50 dB in decades (0 dB≡775 mV≡1 mW into 600 Ω)

Accuracy:
Voltage ranges: ±(2% of f.s.d. + 0.5% of measured value)
Current ranges: ±(2% of f.s.d. + 1.5% of measured value)
Resistance ranges: all ranges from 10 Ω to 1 MΩ:
                            ±[2%(2+x+1 / x)+0.4%(1+x)]
                        in 3 Ω range:
                            ±[2%(2+x+1 / x)+1.3%(1+x)]
                        (x = centre-scale/measured value)
dB ranges: ±(0.18 f.s.d. + 0.002 dB measured value)

Power supply:
Supply voltage: 6 V (4 penlight batteries)
Current consumption: 25 mA
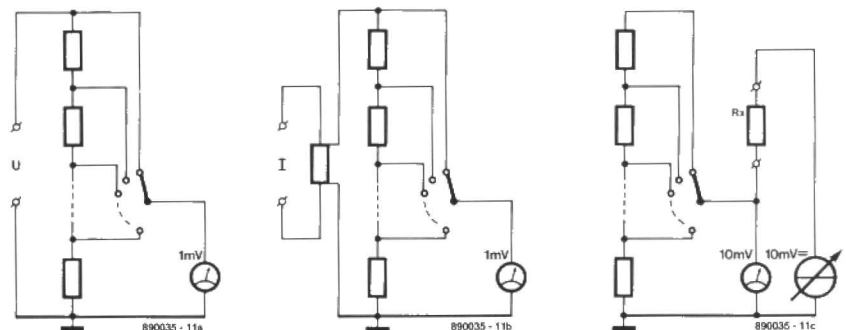
Fig. 1.   Basic measurement techniques applied in the analogue multimeter: voltage (a); current (b) and resistance (c).

voltage divider is determined almost exclusively by the tolerance of the resistors, since the error introduced by the ratio of their values alone is of the order of a few hundredths of a per cent. This error is, however, slightly higher in the 100 V, 300 V, 10 Ω and 3 Ω ranges because the values of R_{12} and $R_{13} \| R_{14}$ are then so small that the resistance of the solder connection becomes significant. There is, therefore, little point in using 0.1% resistors for R_{13} and R_{14}. None the

less, the deviation caused by the bottom resistor in the ladder is made as small as possible by connecting R_{13} and R_{14} in parallel and so ruling out additional deviation caused by an incorrect nominal value.

Another composite resistor, R_1-R_2, is seen at the top of the ladder network. In principle, a single 681 kΩ resistor would have been all right here, but the maximum permissible input voltage of the multimeter, 300 V_{rms}, results in a peak

value across (R_1+R_2) of $200\sqrt{2}\cong282$ V. This is close to, and in some cases over, the maximum voltage that may exist across most types of 0.25 W or 0.33 W resistor. The required 'high-voltage' resistor is, therefore, made from two series-connected resistors with about equal values and a total equivalent value of 681 kΩ. In the worst case, each of these resistors thus carries only about half the maximum anticipated voltage. Input coupling capacitor C_1 serves to
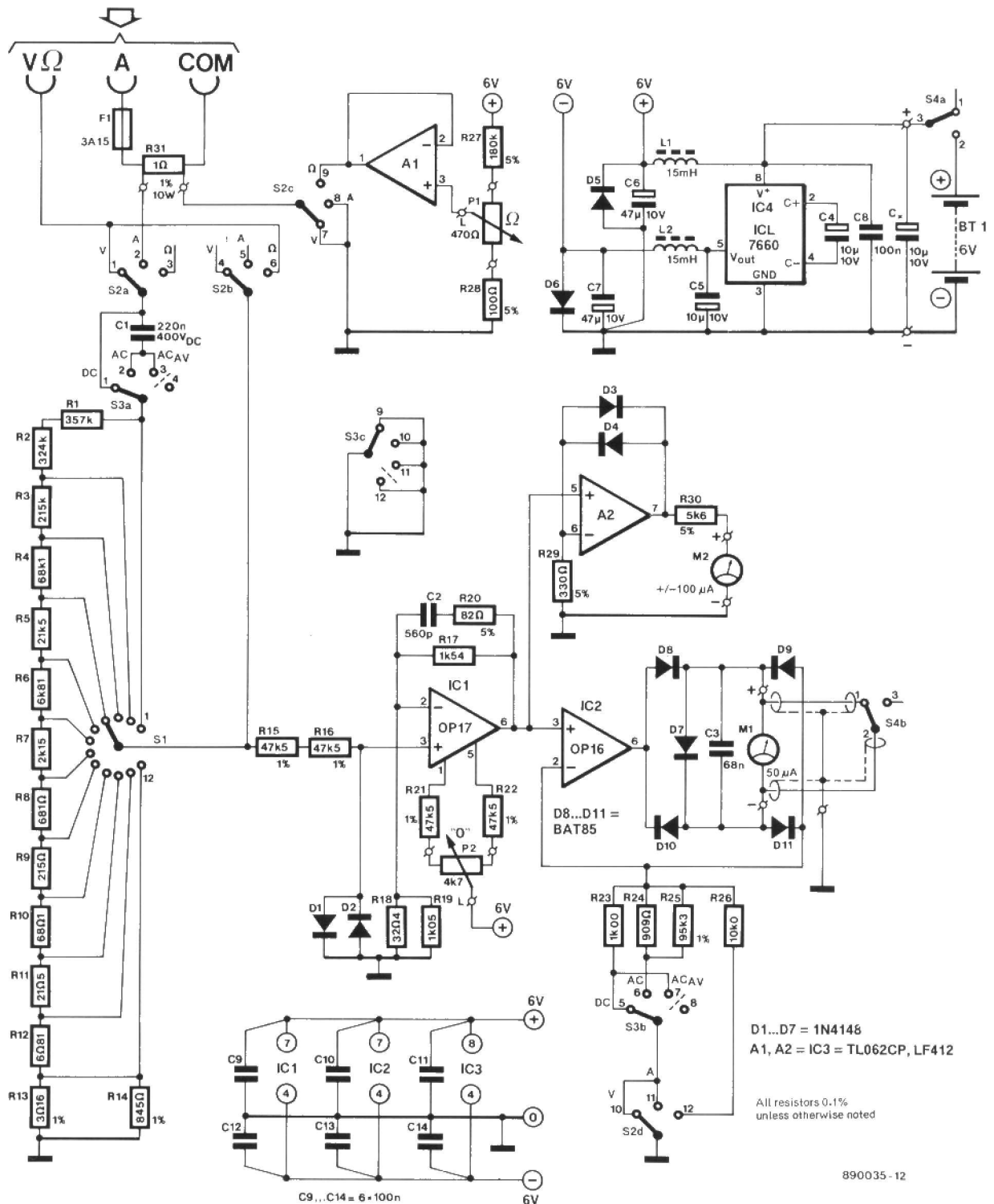
Fig. 2. Circuit diagram of the analogue multimeter.

remove DC components when alternating voltages are being measured. The alternating voltage across C_1 is negligible at most frequencies, and the direct voltage across it is 300 V at the most. The capacitor drops some alternating voltage at input frequencies below 10 Hz only. Fortunately, a measured alternating voltage of this or a lower frequency will rarely have an amplitude that causes the 400 VDC specification of C_1 to be approached or exceeded.

The voltage source of Fig. 1c is found back in the circuit diagram as sub-circuit A_1-P_1-R_{27}-R_{28}. Preset P_1 allows the output voltage of A_1 to be adjusted such that the instrument indicates 0 Ω with short-circuited input terminals. Fluctuations on the battery supply voltage are thus compensated. The 0 Ω adjustment also provides a convenient 'battery-low' indication: as soon as P_1 has to be turned almost fully counter-clockwise to achieve the 0 Ω indication, it is time to replace the battery pack.

Power supply

The heart of the power supply is formed by voltage inverter IC_4. The Type ICL7660 comprises an oscillator and a voltage doubler which are used here to generate a negative supply voltage to obviate a second set of batteries. Ripple on the negative output voltage is suppressed by networks L_1-C_6 and C_5-L_2-C_7. The operation of these filters is backed by the special lay-out of the printed circuit board. Diodes D_5 and D_6 protect the operational amplifiers in the meter circuit from permanent damage when their symmetrical supply voltages do not rise at the same rate when the instrument is switched on with S_{4a}.

Input amplifier

The function of the millivolt amplifier is the conversion of the voltage at junction S_1-S_{2b}-R_{15} into a proportional current flow through moving-coil meter M_1. This is done in two steps by two sub-circuits: an input amplifier and a V-I converter.

The input stage around IC_1 is a non-inverting amplifier with a voltage gain of 50. The accuracy of this amplifier is 0.2% within the nominal bandwidth when 0.1% tolerance resistors are used in positions R_{17}-R_{18}-R_{19}. Deviations become greater, of course, outside the amplifier's pass-band: at the 3 dB roll-off frequency, for instance, the voltage reduction is 29%. *Ergo*, the input amplifier must be dimensioned for a bandwidth much greater than that strictly required for reasonably accurate measurements. Allowing an error of 1% (0.08 dB), for instance, means that an amplifier bandwidth is required of about ten times the −1% frequency. A 1%-bandwidth of 20 kHz then corresponds to a −3 dB-bandwidth of 200 kHz. In practice, the bandwidth has been reduced here to about 185 kHz by the addition of C_2.

Potentiometer P_2 compensates off-set introduced by the relatively high amplification of IC_1, and the fact that this is set up as a direct-voltage amplifier. Series resistors R_{21} and R_{22} are added to narrow the otherwise unwieldy range of P_2. Production tolerances on the resistors and the operational amplifier may, however, call for the control range to be shifted a little until minimum off-set is achieved with the potentiometer set to about the centre of its travel. The control range is shifted empirically by correcting the values of R_{21} and R_{22}: a lower value for one necessitates a higher value for the other until minimum off-set is achieved with P_2 set to the centre of its travel. Constructors less inclined to experimenting with this adjustment may replace P_2 and the series resistors with a single 100 kΩ multi-turn preset.

The off-set compensation alone is not sufficient to guarantee stable operation of the sensitive input amplifier: the opamp itself, IC_1, must be a type with low off-set drift to rule out the need for frequent corrections in the setting of P_2. The Type OP-17 from Precision Monolithics used in position IC_1 guarantees very low drift over a wide temperature range. The operating temperature of the device should be allowed to settle, however, for a minute or so after switching on the multimeter.

Overload protection

The input amplifier is protected against overloading by network R_{15}-R_{16}-D_1-D_2. The diodes make it impossible for the amplifier input voltage to exceed +0.6 V or −0.6 V. Any excess input voltage is shunted away via R_{15} and R_{16}. The diodes form a high resistance, and R_{15} and R_{16} have no effect, as long as the input voltage remains within the nominal range. This means that the protection circuit does not normally affect the accuracy of the measurement. It does, however, affect the termination resistance of pin 3 of the opamp, which is sensitive to noise because of the high input impedance. The sensitivity to noise is, in fact, so high that the PCB track carrying the output signal of IC_1 is carefully screened from the input circuitry to prevent oscillation. Recalling that the input amplifier works with very small signals (max. 1 mV), it will not come as a surprise that the carefully designed lay-out of the ready-made printed-circuit board is essential for correct operation of the multimeter. The excellent stability of the analogue multimeter is a feather in the cap of the *Elektor Electronics* PCB design staff.

It should be noted that the high impedance between input terminal and + input of IC_1 causes the capacitance of D_1 and D_2 to become significant in the 3 mV and 3 mA ranges and, therefore, to reduce
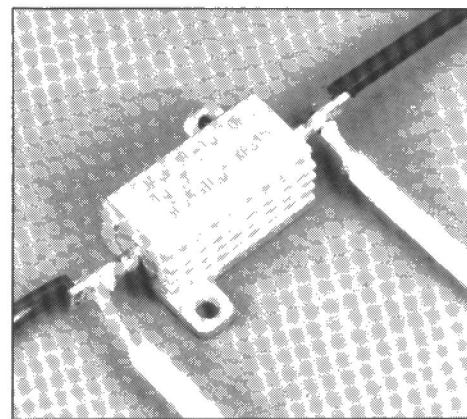
Fig. 3. How to make a shunt from a conventional power resistor.

the amplifier's bandwidth to some extent.

V-I converter/meter driver

The amplified output signal supplied by IC_1 is fed to two sub-circuits. The first of these to be discussed is a precision V-I converter and associated meter driver set up around opamp IC_2.

The moving-coil meter, M_1, is current-driven to achieve maximum accuracy. The meter current, I_m, depends on two factors only: input voltage U_i available at pin 3 of IC_2, and the resistance between the − input of IC_2 and ground (R_{23}, R_{24}||R_{25} or R_{26}). For the switch position shown in the circuit diagram, and assuming that IC_2 provides very high gain:

$$I_m = U_i / R_{23}$$

showing that the accuracy of the voltage-to-current conversion depends on the accuracy of R_{23} only.

Moving-coil meter M_1 forms part of a diode bridge circuit that rectifies the meter current. This results in M_1 indicating the *average value of the rectified current*, whether this is generated by a direct or alternating voltage or current, and irrespective of the polarity of the direct voltage or current. Schottky diodes are used in the rectifier because they have a lower on-resistance than silicon types. This allows the gain of IC_2 to be reduced slightly, resulting in a higher attainable bandwidth.

The bandwidth of the V-I converter is a function of the measured value. The reason for this lies in the fact that the internal resistance of the diodes in the bridge is a function of the forward current, which is, in turn, a function of the measured value. Fortunately, the effects of this annoying variable are practically unnoticable by virtue of a suitably defined bandwidth of the input amplifier around IC_1.

The V-I converter around IC_2 is completed with three switch contacts, a meter overload protection, D_7, and a stabilization capacitor, C_3. Switch contact S_{4b} short-circuits the meter coil when the instrument is turned off. This helps to
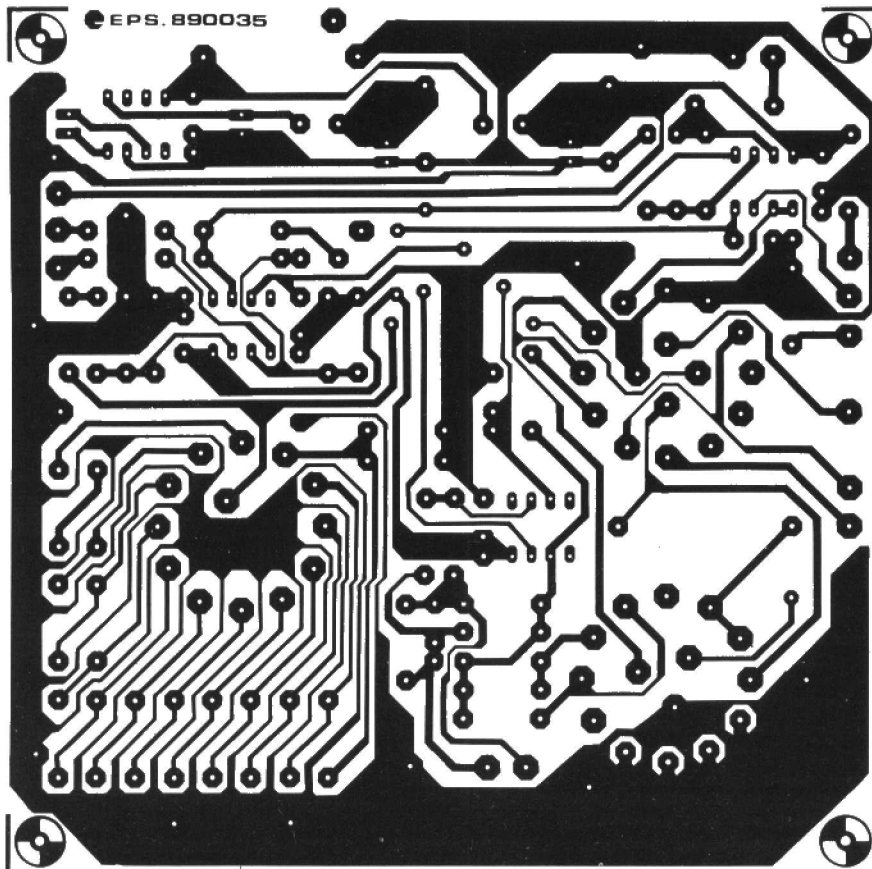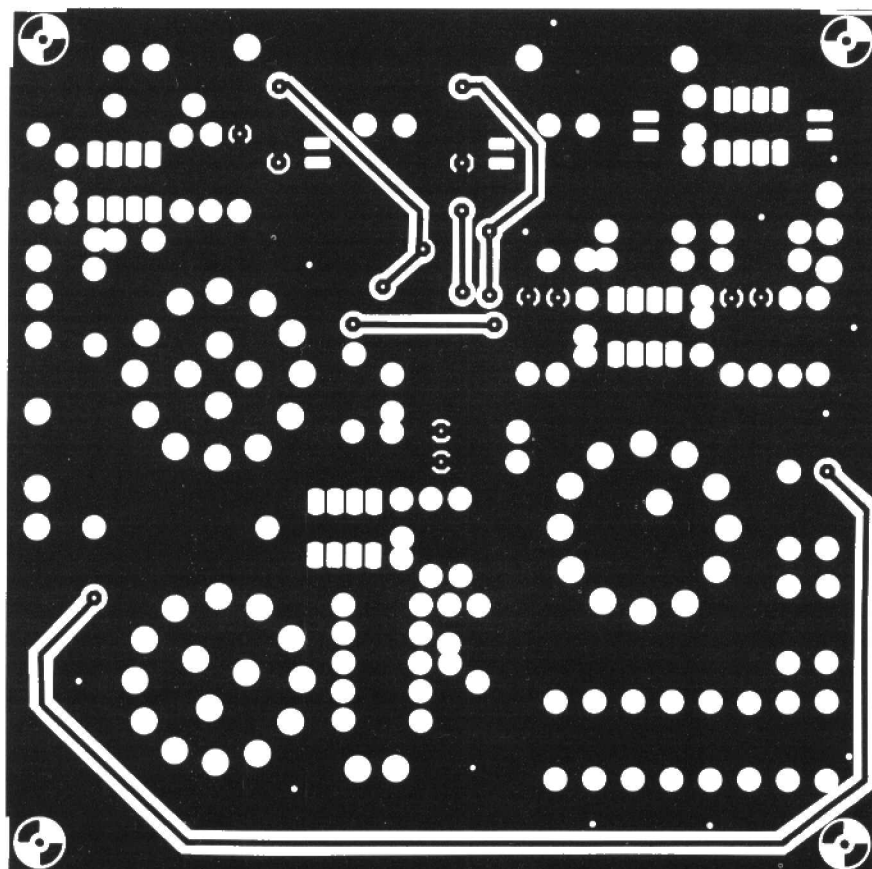
Fig. 4a. Lay-out of the component side (above) and the solder side (below) of the PCB for the multimeter.

Parts list

Resistors:
(all 0.1% tolerance, E-96 series*, unless otherwise noted)

R1=357K
R2=324K
R3=215K
R4=68K1
R5=21K5
R6=6K81
R7=2K15
R8=681R
R9=215R
R10=68R1
R11=21R5
R12=6R81
R13=3R16; 1 %
R14=845R; 1 %
R15;R16;R21;R22=47K5; 1%
R17=1K54
R18=32R4
R19=1K05
R20=82R; 5 %
R23=1K00
R24=909R
R25=95K3; 1 %
R26=10K0
R27=180K; 5%
R28=100R; 5%
R29=330R; 5%
R30=5K6; 5%
R31=1R0; 10 W; 1%
P1=470R linear potentiometer.
P2=4K7 linear potentiometer.

* ElectroMail, telephone (0536) 204555.

Capacitors:
C1=220n; 400 VDC
C2=560p
C3=68n
C4;C5=10µ; 10 V; radial
C6;C7=47µ; 10 V; tantalum
C8;C9;C10;C11;C12;C13;C14=100n

Semiconductors:
D1;D2;D3;D4;D5;D6;D7=1N4148
D8;D9;D10;D11=BAT85
IC1=OP-17 (PMI)
IC2=OP-16 (PMI)
IC3=TL062 (Texas Instruments)
IC4=ICL7660 (Intersil)

all integrated circuits in 8-way dual-in-line (DIL) enclosure.

Miscellaneous:
L1;L2= 15 mH inductor with ferrite core (Ri<25 Ω).
S1= single-pole 12-way rotary switch for PCB mounting.
S2= 4-pole 3-way rotary switch for PCB mounting.
S3= 3-pole 4-way rotary switch for PCB mounting.
S4= double-pole toggle switch.
B1= 4 penlight batteries plus holder.
M1= 50 µA moving-coil meter; Ri<5 kΩ; deflection angle 86°; dimensions: 110×83 mm (e.g. Monacor Type E-11B; order no. 29.0200).
M2= centre-zero moving-coil meter ±100 µA; dimensions: 14×35 mm
F1= fuse 3.15 A delayed action; with panel-mount holder.
PCB Type 890035 (double-sided; not through-plated; available ready-made through the Readers' Services).
Front-panel foil Type 890035-F (available ready-made through the Readers' Services).

Fig. 4b. Component mounting plan. The dashed lines indicate the positions of the screens.

Construction

As already stated, the printed-circuit board designed for the multimeter is essential for correct operation. It is, therefore, not recommended to build the circuit on stripboard or prototyping board. Also, when you plan to etch your own PCB, remember that the quality of the board supplied through the Readers' Services is hard to match with simple materials and tools.

The PCB is double-sided but not through-plated. Screening is fitted as shown on the overlay in Fig. 4b. Commence the construction with making the 9 through-connections on the board. Then fit the power supply parts and check the presence of the correct supply voltages at a number of relevant points on the board.

Proceed with the mounting of the resistors in the input attenuator. The resistors should be fitted about 2 mm above the board surface to minimize stray capacitance. Resistors R_{15} and R_{16} are mounted likewise for reasons of safety.

Circuit IC_1 must be soldered direct on to the PCB: do not use an IC socket! For the other ICs, sockets may be used.

The screening boxes may be installed when all components have been fitted. The piece of tin-plate that screens the solder side of the board is the easiest to make. Short-circuits are prevented by spraying the screen with plastic at the PCB side. The plate must be fitted as close as possible to the PCB — the maximum distance is 2 mm — and is, of course, soldered all around to the ground plane.

The screens at the component side of the board are a little more difficult to install. The locations are indicated by the dashed lines on the overlay. The height of the screens equals that of the bodies of the switches (that is, excluding the threaded shaft). All compartments are, finally, closed with a lid that must, however, be simple to remove.

The components not accommodated on the PCB are connected to it with screened cable. The cable screening is soldered to the ground plane or the tin plates only: at the other side of the cable, it is not connected. The shunt circuit forms an exception: the wires between terminal A to the fuse, between the fuse and the shunt, and between the shunt and the COM terminal should have a cross-sectional area of at least 1.5 mm^2 because they must be capable of carrying currents up to 3 A. The sense connections from the shunt are, of course, made in screened cable (see Fig. 3).

The enclosure is given an attractive appearance with the front-panel foil and the meter scale shown in Fig. 10. The completed scale is carefully cut out and fitted in the meter.

The scale of the polarity indicator must

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protect the coil and the needle from taking damage during transport.

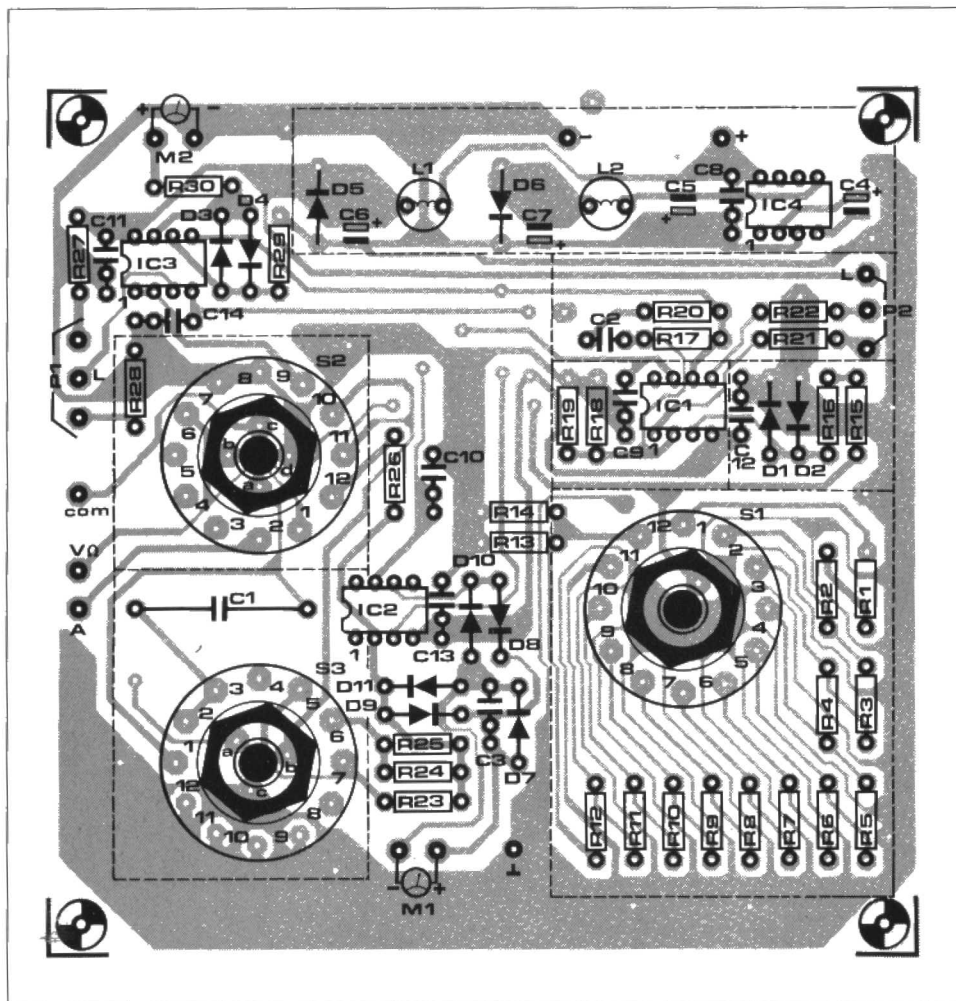The other two switch contacts, S_{2d} and S_{3b}, set the ratio I_m / U_i in accordance with the selected measurement. Resistor R_{23} is used with DC and AC$_{AV}$ measurements. Measurement of sinusoidal alternating voltages require the meter to indicate a 1.11 times higher value. This is achieved by selecting $R_{24} \| R_{25}$, whose equivalent value is 1.11 times smaller than that of R_{23}. As already noted, the sensitivity of the meter must be reduced to 10 mV for resistance measurements. This is accomplished by selecting R_{26} with switch contact S_{2d}.

Indicator circuit

The second signal path from IC_1 leads to an indicator circuit around opamp A_2. This drives a small centre-zero moving-coil meter that indicates the polarity of the measured signal. When M_2 deflects to the right, the V-Ω or A input terminal is positive with respect to the COM terminal. Similarly, when M_2 deflects to the left, the V-Ω or A input terminal is negative with respect to the COM terminal. Meter M_2 does not deflect when alternating voltages and currents are measured. Opamp A_2 is set up as a non-inverting amplifier whose gain depends on the level of the input signal. This is achieved with diodes D_3-D_4, whose equivalent re-

sistance is a function of forward current. The gain of A_2 is inversely related to the input signal amplitude. This results in higher sensitivity around the centre-zero point of the meter, so that the polarity of relatively small signals is indicated reliably. It should be noted that M_2 is merely an *indicator* — off-set compensation with P_2 should, therefore, always be carried out for the main read-out, meter M_1. The response of M_2 around the centre-zero point may be improved by replacing the TL062 with a LF412 which has a lower off-set and off-set drift, but a slightly higher current consumption (5 mA more than the TL062).

Shunt resistor

Resistor R_{31} is the shunt for converting measured current into a proportional voltage. The resistor used is a 10 W type with 1% tolerance. It is not, strictly speaking, a shunt resistor because this normally has four terminals. The two extra wires are created by soldering them to the existing leads where they leave the resistor body. This creates a shunt with sense wires, exactly as indicated by the relevant symbol in the circuit diagram. Fuse F_1 protects R_{31} from destruction when a too large current is passed. It also prevents the value of R_{31} changing gradually as a result of occasional, light, overloads.

be replaced with a home-made one: cut a piece of paper or light cardboard to size and provide on this the marks − (to the left of the scale), + (to the right of the scale) and ~ at the centre (do not use a 0 here because the meter is incapable of indicating the true zero-point).

Practical use

Although the day-to-day use and operation of the analogue multimeter is straightforward and similar to other meters of this type, it is, none the less,

worth wile to note a few particulars.

After switching on, allow a minute or so for the instrument to warm up. The offset-compensation potentiometer is then adjusted for minimum meter deflection, which is not necessarily the same as moving the needle to the 0-indication. Only then may the moving coil meter be nulled mechanically with the plastic screw provided for this purpose. Electronic nulling should always be carried out by indication of the main meter, *not* the polarity indicator.

The multimeter is protected against overloading in the DC/AC/AC$_{AV}$ ranges, but not in the Ω ranges. Do not, therefore, use these longer than strictly necessary, and make a habit of checking the selected range and quantity before making any measurement.

The DC/AC/AC$_{AV}$ selector should be operated with due care. The DC position allows measuring both direct and alternating voltages and currents, or direct voltages superimposed on an alternating voltage. The important thing to remember in all cases is that the meter indicates
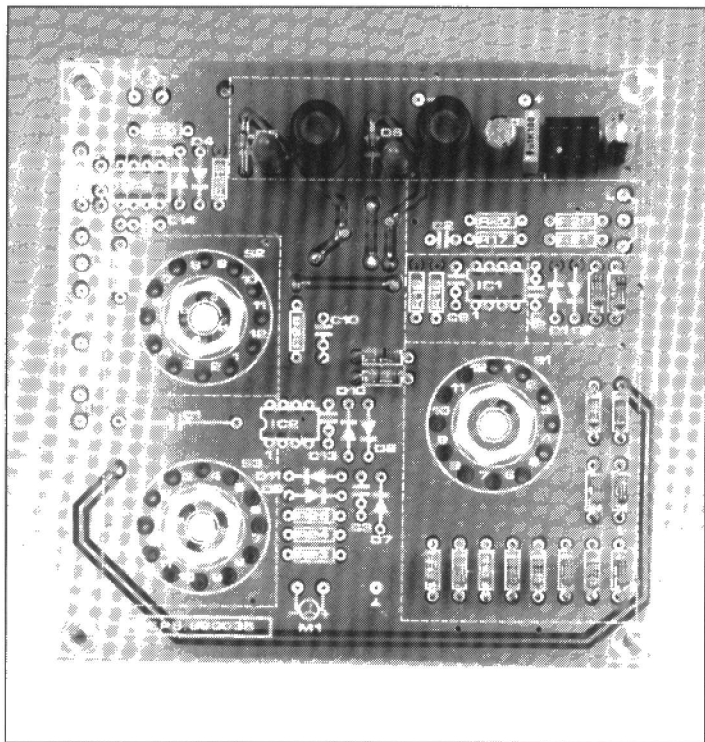

Fig. 5. The first phase of the construction has been completed: the power supply and voltage divider parts are installed on the board.

Fig. 6. Phase two: ready for fitting the screens...

Fig. 7. Phase three: ready for mounting the top plates on the screens...

Fig. 8. The top plates must not be omitted: the front-panel is too far away to achieve proper screening otherwise.

front-panel foil for the multimeter is based on the Monacor moving-coil meter stated in the Parts list. This meter has an indication angle, φ, of 86°. Other meters may be used provided these are capable of covering an angle of 86°. Also, the gain of IC1 may have to be adjusted by changing the equivalent resistance of parallel combination R_{18}-R_{19} according to

$$R_{18}\|R_{19} = R_{17}\varphi / [50(86-\varphi)]$$

When selecting resistors for R_{18} and R_{19} on the basis of the calculated value, it is best to decide on one with a slightly higher value than calculated, and select the other for the best possible approximation of the required equivalent value. The actual value so obtained must be within 0.1% of the calculated value.

How accurate?

The way the analogue multimeter is specified in the shaded box on the first page of this article is perhaps not too common, but not unusual. The data given are presented in a way analogous to that commonly used for specifying the performance of a digital multimeter. The digits specification is, however, replaced by a corresponding percentage of the full-scale deflection (% f.s.d.).

The error calculation applicable to the resistance ranges is rather elaborate and therefore often omitted altogether in manuals supplied with analogue multimeters. A brief description is, however, given here.

The centre of the scale is taken as a convenient starting point for the error calculation because the extremes (nought to the left and infinite to the right) are difficult to work with mathematically. This is why the centre-scale resistance value, rather than the more usual scale multiplier, is printed near the range selector.

Even a brief analysis of the way the meter error is introduced already leads to a rule of thumb. The measurement principle adopted here results in a non-linear scale, and the minimum measurement error is four times the class of the moving-coil meter plus the error caused by the measurement circuit. A class-2 moving-coil meter can not, therefore, measure resistance with an accuracy better than 8%. This minimum value is commonly found as the only specification of many analogue multimeters. However, the error increases exponentially to the left and the right of the centre indication. Fortunately, the meter described here has no fewer than 12 resistance ranges, so that a reading as close as possible to the centre indication is nearly always simple to achieve by operating the range switch.

Fig. 9.   This front-panel foil is available ready-made through the Readers' services. The meter scale is cut out for mounting inside the moving-coil meter (foil shown 76% of true size).

the average value of the rectified voltage.

The AC position enables measuring the alternating component only. Any bias or superimposed DC levels are blocked by a capacitor. The meter shows the average value of the full-wave rectified voltage multiplied by 1.11. This means that the AC position allows reading the effective (r.m.s.) value of sinusoidal input voltages. The AC$_{AV}$ position is used for non-sinusoidal voltages. It is essentially equal to the AC position without the factor 1.11, so that the average value is indicated.

It is often necessary to check whether a direct voltage measurement is valid or not because there may be an alternating component as, for instance, in the case of an oscillating amplifier. When an alternating component is measured, this may invalidate the measured direct voltage. This need not be so in all cases,

however, since the correctness of the direct voltage indication depends on the ratio of the direct voltage to the peak value of the alternating voltage. The previously taken DC measurement is incorrect if this peak value is greater than the direct voltage.

It is possible to use the multimeter as an ammeter and a voltage meter simultaneously, but only if one terminal can be connected to a common junction in the circuit under test. This is the case in, for instance, resistance measurement to the current-voltage method. Current and voltage can then be measured separately by switching between A and V with S_2. This useful feature is available because the shunt circuit is not broken, and the VΩ terminal is not disconnected.

Moving-coil meter

The design of the scale supplied with the

SCIENCE & TECHNOLOGY

Forecasting flickers in the field

by Dr Bill Stuart, Geomagnetism Research Group, British Geological Survey, Edinburgh

Rapid and hitherto unpredictable fluctuations in the Earth's magnetic field have for centuries been a problem to navigators and, in more recent times, for operators of cable, radio and radar systems. Spacecraft become subject to changes in atmospheric drag that may affect their orbits and re-entry glide paths by many tens of kilometres. Automation of geomagnetic data collection and processing has now made immediate information about the field available to laboratories on a national scale and will make it accessible to professional users internationally. A global mathematical model enables field values to be calculated with good accuracy for any part of the world and gives forecasts for some years ahead.

"I saw an aurora to the South and noted simultaneously a great movement of the magnetic needle. When I announced this to Professor Celsius he said he too had noticed such a disturbance but had not mentioned it in order to see whether I too would light on the same speculation. Professor Celsius had, by letter, requested Mr Graham in London to observe his own needle for several days. The magnetic needle at London had just such an unusual motion at the same time as here at Upsala."

That was how, on 1 March 1741, a student of Andreas Celsius confirmed observations made some 20 years earlier by George Graham that a finely pivoted magnetic compass was subject to sudden, unexplained disturbances, and added that the disturbances occurred simultaneously in London and Upsala and that they related, in some way, to displays of bright aurora. On 20 October 1746, Celsius measured a disturbance of 4°10' of arc in the compass. (On 7, 8 and 9 February 1986 compass deviations reached over 10° during a severe magnetic storm that lasted for 42 hours.)

A century later Baron Alexander von Humbolt wrote letters to the Royal Society (23 April 1836) and to the Royal Geographical Society (10 January 1838) proposing that Great Britain, with its worldwide colonial influence, should join in a scheme to establish geomagnetic observatories using the standardised instruments and measuring techniques devised by Carl Friedrich Gauss, and participate in a worldwide study of geomagnetic disturbance. He worte: "I think the subject is not without importance to seamen."

Record prize

By 1841, 50 geomagnetic observatories had been established where compass needles were observed by hand and eye once an hour or sometimes every five minutes. The Admiralty offered a prize of £500 for the invention of magnetic self-recording arrangements. It was won by Charles Brooke, who used a light beam reflected from a mirror attached to the end of the compass to magnify the compass movements, and recorded the beam's motion on a sheet of photographic paper attached to a slowly rotating drum. Co-ordinated measurement of the Earth's magnetic field throughout the world had begun.

Graham and Celsius used only a sensitive compass to detect changes in the direction (D) of the geomagnetic field in the horizontal plane. Gauss devised suspensions that responded to changes in the strength of the horizontal component (H) and the strength of the vertical component (Z); he used a torsion suspension and counterbalanced pivot respectively. In that way variations of the full vector field could be recorded. Such variometers record only field changes about an arbitrary zero, or baseline. To convert their records to actual field values, absolute measurements must be made and compared with the variometer record to calibrate its scale and ascribe an actual value to the baseline. Gauss introduced this system of operating and devised techniques for making absolute measurements using standard magnets to create deflections at fixed distances and for measuring their oscillation period.

Although very effective, such essentially anologue instruments are not compatible with digital logging systems and computer control. An electronic magnetometer, known as a fluxgate, was developed in World War II for airborne submarine detection and is still an essential part of *Nimrod's* airborne tactical system. It has been adapted for use in geomagnetism. In the instrument, an alternating electrical current in a winding around a core of ferromagnetic material magnetises the core to saturation at each half cycle. An ambient magnetic field along the core produces asymmetry in the cycle of magnetisation and causes second harmonic distortion in the voltage induced in a secondary winding around the core. The amplitude and phase of the second harmonic is a measure of the magnitude and sense of the ambient field.

Because a fluxgate sensor is a single-axis device, sensors arranged at right-angles are used to obtain the geomagnetic vector. The inherent instability of electronic components used for amplification and rectification in the instrument means that it, too, has to be calibrated against precise absolute measurements from time to time.

Proton magnetometers

Modern absolute measurements make use of the fact that atomic particles have a magnetic moment, by virtue of their electrical charge and their spin or orbital angular momentum, which causes them to precess in a magnetic field in an analogous way to that of a gyro under the influence of gravity. Normally, individual precessions are randomly distributed in phase and cannot be detected. But if a very strong magnetic field is applied in a direction perpendicular to the ambient field the precession takes place about the axis of the field and the atomic particles become, in effect, aligned in that direction. If the added field is removed suddenly the particles begin to precess about the remaining ambient field from that direction and are, therefore, in phase coherence; that is, they all do it in unison. Their precession can be detected by the voltage their combined magnetic moments induce in a coil around the sample. Its frequency is a measure of the ambient field. Water, or a hydrocarbon, provides a convenient source of protons in the hydrogen nuclei, and proton precession magnetometers have become the standard for geomagnetic field measurement because of the precision (one or two parts in 10^6) to which their fundamental constants are known, and because they are portable and easy to operate.

Proton magnetometers measure the strength of the field without providing information about its direction. For measuring directional components of the field they are adapted by applying control fields, generated in what are called Helmholtz coils, arranged to cancel out all but the component which is to be measured. The simple example is a coil assembly with an axis arranged horizontally in the geomagnetic meridian. The generated field is adjusted to cancel H, leaving Z to be measured.

Core of the earth

The main geomagnetic field is caused by electrical currents flowing in the fluid part of the Earth's core, driven by mechanical forces from the Earth's rotation, the heat and pressure of depth, radioactive decay and by chemical changes taking place in the core itself. The main field changes continuously with time. For example, the direction of magnetic North in Britain moves eastward at a rate of one degree every five years, and the strength of the main field decreases by one per cent every ten years. One reason for this so-called secular change is that the core fluid is a perfect electrical conductor and lines of magnetic field which cross its surface are frozen in relation to the fluid. So the movement of field lines measured on the Earth's surface is a direct representation of the motion of the surface fluid of the core. Recent studies reveal that abrupt changes occur in the rate of secular change, caused perhaps by inner processes of mountain building within the core itself.

One task of co-ordinated geomagnetic measurement is to chart the main field and secular change for studies of the Earth's deep interior. It is also necessary for navigation, which uses geomagnetic reference as much now as at any time in the past. Modern navigational charts are derived from a mathematical model which fits a dipole, quadrupole and higher magnetic multipoles to the global set of measurements. Such mathematical models are ideal for incorporation in flight-deck computers of aircraft and missile guidance systems.

The main field extends outward into space where it comes under the influence of the solar wind, a stream of protons and electrons boiling off from the Sun and flowing into interplanetary space. The geomagnetic field creates a bubble in the solar wind that is called the magnetosphere, preventing it from reaching Earth directly. In the magnetosphere, the geomagnetic field is compressed on the sunward side and drawn out in an extended tail on the nightside in a dynamic state of balance with the forces of the solar wind. The shape of the magnetosphere and its balance are maintained by electrical currents flowing in a well-defined network of circuits within the magnetosphere and in the ionosphere. It is the magnetic effect of these currents which creates geomagnetic disturbance. Because there is a great deal of variability in the solar wind it is very blustery, so geomagnetic activity is erratic and sometimes extremely intense.

Automatic techniques

In 1987 the three standard geomagnetic observatories in Britain, now operated by the Geomagnetism Research Group (GRG) of the British Geological Survey (BGS), finally dispensed with photographic recording of geomagnetic activity and adopted automatic methods controlled by a central computer in Edinburgh.

A microcomputer at each observatory is programmed to sample the field experienced by each fluxgate sensor every 10 seconds, to transfer these data to store and to perform routine computations on them. Minute values are computed using a filter to exclude very high frequency field fluctuations from the final data set. Mean hourly values are computed at the end of every hour. The range of disturbance (maximum value minus minimum value) for each component of the field is also computed, at the end of each hour, for the preceding hour and three hours in conformity with international convention. These ranges are converted into a standard index of magnetic disturbance, known as K values, which is compatible with those derived from handscaling the old magnetograms. When the processed data are transferred to Edinburgh an analogue record is plotted, corresponding with the old photographic daily magnetogram.

All the prepared information is available for use a fraction of a second after the end of each hour. In practice it is stored on cartridge tape and awaits instructions from the computer in BGS's Edinburgh office. The Edinburgh computer is connected to each observatory by the public telephone network using auto-dial modems. The eventual aim is for the central computer to call up each observatory on its own initiative and retrieve and process the data into the final forms that users need. This means that self-checking procedures have to be developed to replace the manual tests to maintain veracity and quality control. These tests, often subjective, are crucial to maintaining the high quality of operation expected of a standard geomag-
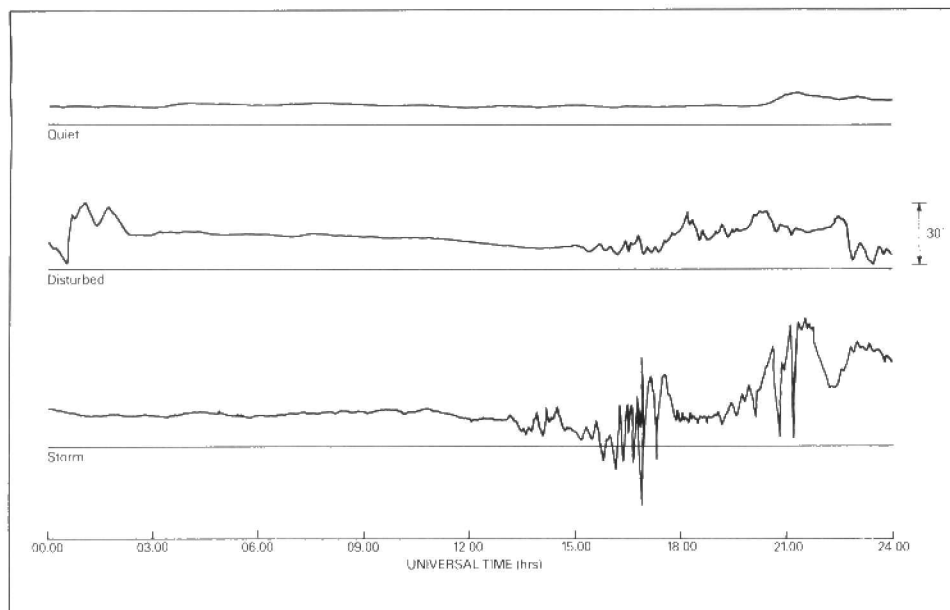
Examples of daily magnetograms showing the most common range of disturbance.

netic observatory. In the meantime, while automated quality control is being tested, the procedure is for operators to call up data every working day so that the processing is supervised.

Getting it right

The most important quality control measure is to compare the fluxgate readings with precise absolute measurements. Most magnetometers, fluxgates and suspended magnets are subject to drifts of calibration and baseline. In classical observatories absolute measurements are made weekly and variometer baselines are adjusted to take account of the differences between variometer readings and measured values. Manual absolute measurements take a relatively long time, some 10 to 15 minutes, and on occasions when the geomagnetic field is even only slightly disturbed the errors in timing often exceed the real difference between variometer and absolute value. The automatic system used by BGS uses proton precession magnetometers in carefully aligned coils which produce the control fields for vector information to perform baseline reference measurements (BRMs) every hour in an automatic sequence. The BRMs are synchronised with the minute values from the fluxgates and are compared with them 'on the hour' to produce a continuous baseline for the variometers. BRMs are absolute measurements with but one exception: the coils can become misaligned or the pillars on which they are placed may tilt gradually. In such circumstances the control fields may introduce errors in the absolute value measured. So manual absolute measurements are still needed, but now only two or three times a year to correct for the gradual and very small changes in the BRMs. The synchronisation of BRM comparisons virtually eliminates errors due to geomagnetic disturbance.

Apart from being technically satisfying to those engaged in the work, the development of BGS's Automatic Remote Geomagnetic Observatory System (ARGOS) makes geomagnetic data available to users in science and commerce in a form that can be handled by computer. It does so a great deal more rapidly than did the classical techniques. Photographic charts used to be changed daily: they had to be developed, fixed and dried; manual scalings had to be independently verified and eventually typed into a computer to be put into digital form. At a single observatory that process took a day or two, and when regional or global data are needed the delivery time took weeks, months or even years.

Editing and availability

When the ARGOS data have been called
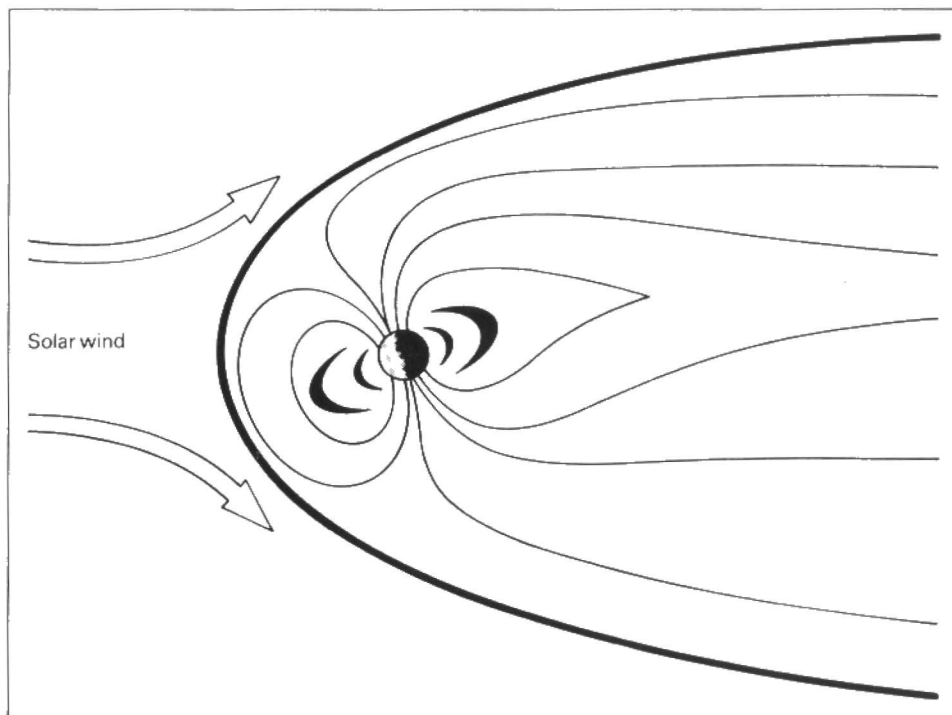
In the magnetosphere, the solar wind is deflected about Earth by the geomagnetic field. On the day-side, solar wind pressure compresses the magnetic field lines while electromagnetic friction drags the night-side field lines into a long tail. Shaded areas near Earth represent the radiation belts where high-energy protons and electrons are trapped in geomagnetic field lines, creating dangerous conditions for astronauts and equipment in spacecraft.
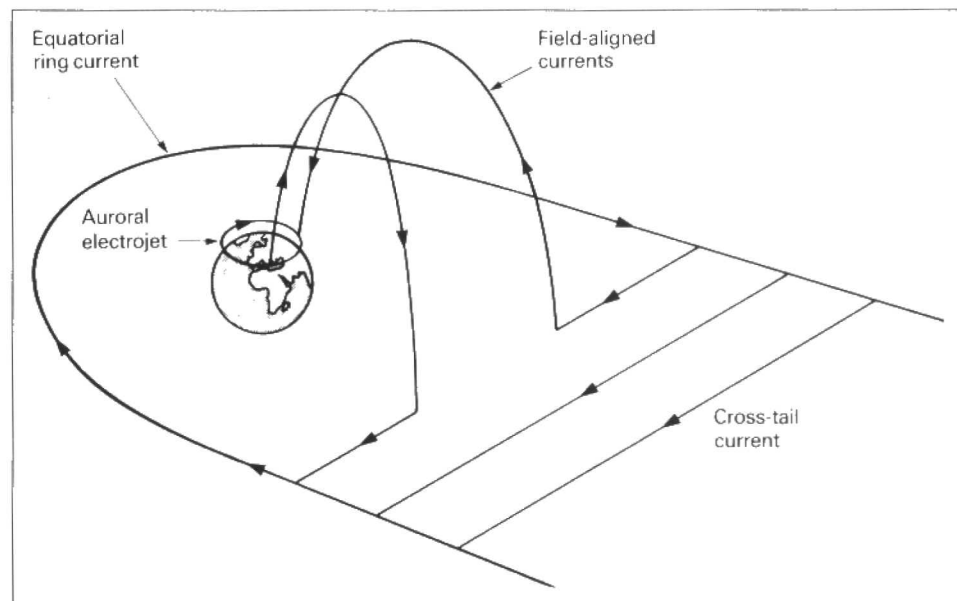
Skeleton representation of electrical currents brought about by the solar wind's deformation of the magnetosphere, maintaining the balance between the pressure of the solar wind and the magnetic pressure of Earth's field. It is the magnetic fields of such currents that cause geomagnetic disturbances.

up to Edinburgh and verified to be correct, they are screened for spikes, which are errors caused by transient interference, and for drop-outs, which are lost data bits. Such errors are almost entirely caused by noise on the telecommunications line. They are edited out and the analogue record is plotted for quick reference. Data are stored for computation of daily, monthly and annual summaries, but are also transferred to files on BGS's mainframe computer where they are available to users, with other data maintained by GRG in its role as a World Data Centre for geomagnet-

ism, through JANET, the Joint Academic Computer Network. JANET is accessible by research workers in all UK universities and research institutes and, via commercial networks, to other users in the UK, Europe and around the world.

The BGS Geomagnetic Information and Forecast Service (GIFS) holds historical data sets from the world's observatories. It also carries simple indicators of solar activity relevant to forecasting geomagnetic disturbance levels, and it provides access to:

(a) Annual mean values of the geomagnetic field components from 511 observatories, dating back to 1813.

(b) Hourly mean values of the three field components for the three UK observatories over the previous 12 months; the file is updated daily.

(c) Hourly and three-hourly values of the range of activity, and the index based on it, for the previous 12 months, also updated daily.

(d) Files of the local activity index for the three UK observatories, extending back to 1940.
Files of planetary activity indices and what is known as the geomagnetic character figure, back to 1932, with files of daily sunspot numbers and solar 10·7 cm radio flux. These historical records will be extended back to about 1850 in the near future.

(e) A program that enables local values of the geomagnetic field components to be calculated from a global field model, known as the International Geomagnetic Reference Field.

(f) A forecast of geomagnetic activity for the next 24 hours and, with less confidence rating, for the following 26 days (the next solar rotation). The forecast includes current and predicted values of the solar 10·7 cm flux.

Forecasting space weather

The power of GIFS lies in combining the historical record of geomagnetic and solar activity with current data and the activity forecast. Forecasting space weather, which is what geomagnetic activity means, was not worth while when the most recent data available were days or weeks old.

A forecast of space weather is extremely valuable for people engaged in high technology because it has to do with how much and what type of particle radiation is reaching Earth's magnetosphere through the solar wind. Its bearing on the magnetosphere includes radiation hazards to astronauts making space walks and electrostatic charging of the spacecraft themselves, which can cause electronic components to burn out or circuits to be falsely triggered, spurious commands and the black-out of communication. Raised ionospheric temperature and density increase atmospheric drag at altitudes from 100 to several thousand kilometres and affect the orbits of low-flying spacecraft used for defence, navigation and commercial purposes. The re-entry glide path of the space shuttle is shortened by 75 km at times of geomagnetic disturbance. Ionospheric temperature may increase by 1000 °C and its density increases by

up to 50 per cent at times of geomagnetic disturbance; these figures correspond with 5 °C and 0·5 per cent respectively for a serious storm in the lower atmosphere. During geomagnetic disturbance ionospheric winds reach many thousands of kilometres per hour. One of the main applications of space science during the next 30 or 40 years will be observation of Earth for economic and ecological planning from long-life low-altitude platforms such as the space stations planned in the *Columbus* programme. Managing such spacecraft depends heavily on knowing flight conditions and having advance warning of major changes.

One important application of knowing space weather lies in monitoring the orbital decay of spacecraft whose life is over. Since 1957 oer 30 000 artificial satellites have been launched for civil and military purposes. All will eventually fall back to Earth, and most do not burn up completely as they re-enter the atmosphere. It is especially important to be able to predict the landfall of those powered by nuclear reactors. This application and techniques for predicting landfall are likely to become more and more important.

Key military targets (weapons silos, command centres and so on) are protected against nuclear attack by especially thick and reinforced concrete. Certain long and intermediate range ballistic missiles are designed to have an accuracy of better than 50 metres to be effective against such 'hardened' targets. The weapons are rendered useless if precise and up-to-the-moment information about the drag and deflection they might experience in falling through the ionosphere is not available.

Defence experts also need space weather information for the sort of telecommunication that depends on the electron density in the ionosphere. Reliable use of 'over-the-horizon' radars depends on knowing the effective heights of the ionospheric layers and their distribution. Geomagnetic disturbance itself causes electromagnetic induction in the ground and in long conductors such as pipelines, telephone or television cables, railway signalling systems and power lines. In Finland recently a magnetic storm generated a transient surge current of 160 amperes in a power line rated at 350 amperes and carrying a standing load of 300 amperes.

None of these needs for information could be satisfied before automatic operation of observatories came into being. Other countries have modernised their geomagnetic observatories, notably Canada which led the way 15 years ago with digital recording, but none has yet taken the important step of full automation with central control and processing. There are about 200 geomagnetic observatories around the world and

more than two-thirds of them still use instruments and methods based on Guass's techniques and Brooke's photorecording of 1841.

To provide global information in the rapid time needed for useful forecasting, BGS and the US Geological Survey (USGS) are promoting a co-operative programme called INTERMAGNET, through which geomagnetic observatories around the world will transmit real-time data by satellite to three or four communication centres which will collect and process the data for users. A pilot scheme is under way, supported by the Air Weather Service of the US Air Force, in which six USGS geomagnetic observatories are being equipped with automatic recording equipment and transmitters. The three UK observatories will have transmitters fitted to ARGOS and receivers will be sited at the USGS headquarters at Golden, Colorado and the BGS observatory at Hartland, England. The two sets of observatories will send data to GOES, a US Department of Commerce satellite, which will beam the information to the two receiving stations. In Britain the data will be transferred from Hartland to Edinburgh by dedicated telephone line and will be made immediately available on GIFS. In the US the information will be relayed directly to the Air Weather Service.

USGS and BGS have invited all the world's observatories to participate in the programme. Already Canada, France, India, Japan, South Africa and the Soviet Union have made firm commitments. It is expected that over 100 observatories using standardised equipment will be linked together by INTERMAGNET by the end of the century. Observatories using classical measuring techniques have to employ many highly-skilled people, so it is hardly surprising that they are found mainly in the more developed countries. There is not enough coverage in the southern hemisphere and almost none in the oceans, and that bias is reflected in field models. Areas that produce no data are modelled by using information from areas where there are plenty of data. INTERMAGNET includes plans to set up unattended instruments in countries whose priorities and resources cannot justify their maintaining a programme of geomagnetic research, and on oceanic islands from which geomagnetic measurements have never before been available. Through this, models of the main field and secular change will become a great deal more accurate and more rapidly available.

PRACTICAL FILTER DESIGN (5)

by H. Baggott

Last month we discussed the design of wide band-pass filters. This month's instalment in the series deals with narrow band-pass network

Not only the gradient of the skirts of a band-pass filter, but also its relative bandwidth, B, is important – see Fig. 26. Steep slopes and a narrow pass band are two conflicting requirements, so that the filter design is really a compromise.
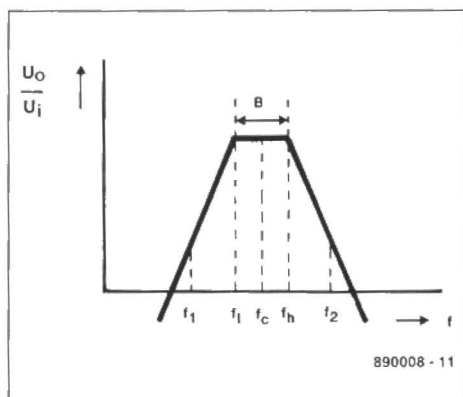
Fig. 26. Characteristic curve of a band-pass filter: the bandwidth, B, is the section between the −3 dB points

The bandwidth of a band-pass filter is closely associated with the Q-factor, which is the ratio between the centre frequency, f_c, of the pass band and the −3 dB bandwidth:

$$B = f_h - f_l \qquad [31]$$

The centre frequency of the filter is not the arithmetic mean of the −3 dB frequencies but is calculated from:

$$f_c = \sqrt{f_l f_h} \, , \qquad [32]$$

If the Q-factor is higher than about 10, for all practical purposes

$$f_c = (f_l + f_h) / 2 \qquad [33]$$

The computation of a band-pass filter should be based on a symmetrical design. If, for instance, a given attenuation is wanted at a number of frequencies that lie outside the pass band (indicated by f_1 and f_2 in Fig. 26), it is necessary to calculate with the aid of the formula $f_c^2 = f_1 f_2$ at which frequency the most stringent requirements occur. All further computations must then take account of these findings.

Assume that we want to design a band-pass filter with a centre frequency of 1000 Hz, a bandwidth of 250 Hz and an attenuation of not less than 40 dB at 400 Hz and 3000 Hz. The frequency corresponding to 400 Hz at the other side of f_c is calculated as follows:

$$400 f_2 = 1000^2, \text{ so that}$$

$$f_2 = 1000^2 / 400 = 2500 \text{ Hz},$$

that is, an attenuation of 40 dB at 400 Hz is already achieved at 2500 Hz. If we choose an attenuation of 40 dB at 3000 Hz, the corresponding frequency at the low-frequency side of the curve will be:

$$f_1 = 1000^2 / 3000 = 333.3 \text{ Hz}.$$

This does not give the required attenuation and, therefore, we must choose the first solution.

The gradient of the slopes of the curve is calculated from the ratio of the bandwidth at two equally attenuated frequencies and the pass band:

$$\text{gradient} = (2500{-}400) / 250 = 5.25.$$

It then becomes a question of looking up the frequency curves that will be published in one of our forthcoming issues and finding a filter that has an attenuation of not less than 40 dB at a normalized frequency of 5.25. This is, of course, the same method as used for low-pass and high-pass filters.

Passive band-pass filters
The first and simplest way of designing a passive band-pass filter is computing

appropriate low-pass and high-pass sections and connecting these in series as described in the previous instalment. In the case of narrow band-pass filters, it is, however, necessary to retain correct impedances at the input and output of the composite network, and this is achieved by inserting a T- or π-attenuator as shown in Fig. 27. The attenuation should be of the order of a few decibels.

A rather more sophisticated band-pass filter may be designed by first computing a low-pass section and then transforming the results. This is rather simpler than it sounds: capacitors are replaced by a parallel network of a capacitor and inductor, while an inductor is replaced by a series combination of an inductor and a capacitor as shown in Fig. 28. The formulas for calculating the components are:

$$L = 1/C(2 \pi f_c)^2 \qquad [34]$$

$$C = 1/L(2 \pi f_c)^2 \qquad [35]$$

An example of a practical design is given in Fig. 29. Here again, we first determine the gradient of the slopes and then compute a low-pass section. The various components are calculated for a low-pass filter with a bandwidth equal to the −3 dB pass band of the band-pass filter. The values are then recalculated with the aid of formulas [34] and [35]. This method is fast and gives good results.

The operation of passive filters is strongly influenced by the Q-factor of the inductors used. The Q(uality) of an inductor is determined by the ratio of its impedance to its d.c. resistance:
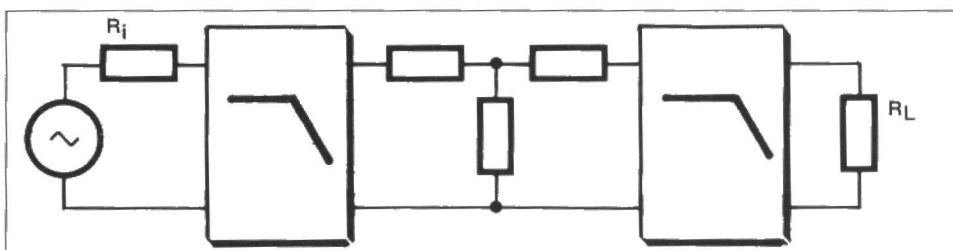
Fig. 27. When two filters are cascaded, it is necessary to insert an impedance-correcting attenuator.
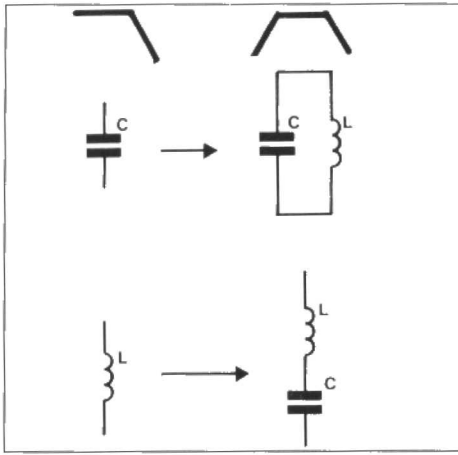
Fig. 28. Illustration of the manner in which the values of components calculated for a low-pass section may be transformed into those required for a band-pass filter.

$$Q = \omega L / R \qquad [36]$$

If care is taken to ensure that the Q-factor of the inductors is several times larger than that of the filter, deviations from the theoretical curves will be minimal.

Active band-pass filters

We shall discuss the design of active band-pass filters on the basis of two fairly simple circuits. The first is a design with multiple feedback paths as shown in Fig. 30. With this design it is possible to start with available (standard) values of capacitors and then calculate the values of the resistors. This is a worthwhile advantage if it is remembered that the computation of a band-pass filter can be pretty complex. Each pole requires a separate filter, so that a design quickly becomes fairly large. A fourth-order filter, for instance, consists of four sections.
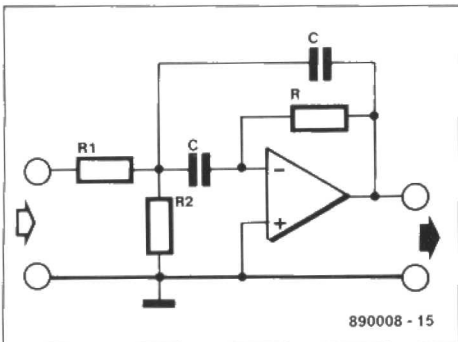
Fig. 30. Computations for an active band-pass filter with multiple feedback paths remain relatively simple. This type of filter may be used for Q values of not greater than 25.

To determine what type of filter and what slope gradient are required for a given application, it is necessary to start with the calculations indicated at the beginning of this instalment. The results will enable the pole characteristics of the wanted filter type to be found in the tables, after which
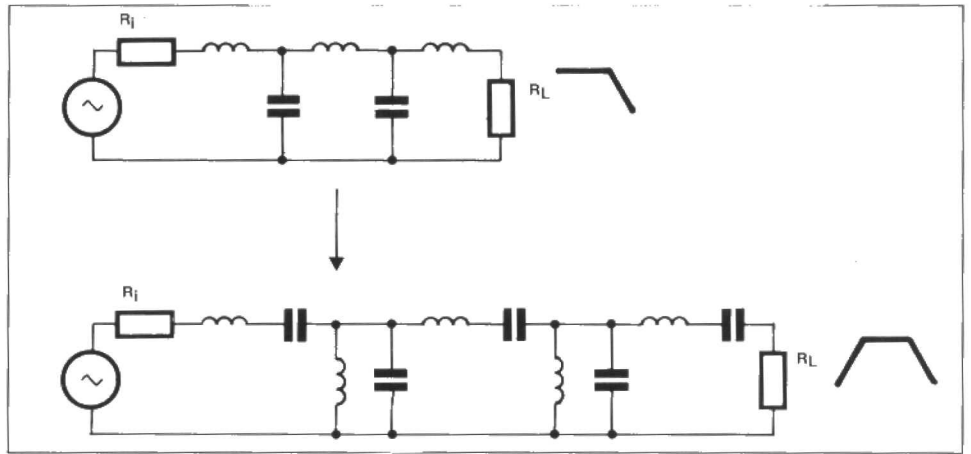
Fig. 29. Practical realization of the transformations illustrated in Fig. 28.

the specific calculations to arrive at the centre frequency and Q of each section can be carried out. In the case of a pair of complex poles this is done as follows:

$$C = \alpha^2 + \beta^2 \qquad [37]$$

$$Q_s = \sqrt{(\{C+4Q^2+Q\sqrt{[(C/Q)^2+16Q^2-8\alpha^2+8\beta^2]}\} / 8\alpha^2)} \qquad [38]$$

$$D = [\alpha Q_s+\sqrt{(\alpha^2 Q_s^2-Q^2)}] / Q \qquad [39]$$

$$f_{sa} = f_c / D \qquad [40]$$

$$f_{sb} = D f_c \qquad [41]$$

where f_{sa} and f_{sb} are the centre frequencies of the two band-pass filters: they have the same value Q_s.

For a real pole, the calculation is somewhat simpler:

$$Q_s = Q / \alpha \qquad [42]$$

The centre frequency of a real-pole section is equal to the central frequency of the entire band-pass filter.

The amplification, A_s, of each stage contained in the filter is:

$$A_s = A\sqrt{[1+Q_s^2(f_s/f_c-f_c/f_s)^2]} \qquad [43]$$

where A is the overall amplification, f_s is the resonant frequency of the section and f_c is the centre frequency of the complete filter. It is prudent to distribute the overall amplification equally over the sections. Once all these computations have been completed, three parameters of each section are known: Q_s, f_s and A_s.

After a suitable value has been chosen for capacitors C, the value of the resistors can be calculated:

$$R_3 = Q_s / \pi f_s C \qquad [44]$$

$$R_1 = R_3 / 2A_s \qquad [45]$$

$$R_2 = R_3 / (4Q_s^2-2A_s) \qquad [46]$$

The amplification, A_s, must not be greater than the amplfication of the multiple-feedback path circuit $2Q_s^2$. At the same time, it should be noted that the open-loop amplification of the opamp must be much greater than $2Q_s^2$.

If Q-factors greater than 25 are required, the design of Fig.31 should be used: it enables Q-factors of up to 100 to be achieved. The component values are calculated with the aid of the following formulas:

$$R_3 = R_4 =1 / 2\pi f_s C \qquad [47]$$

$$R_1 = (2Q_s R_3) / A_s \qquad [48]$$

$$R_2 = R_1 A_s / (2-A_s) \qquad [49]$$
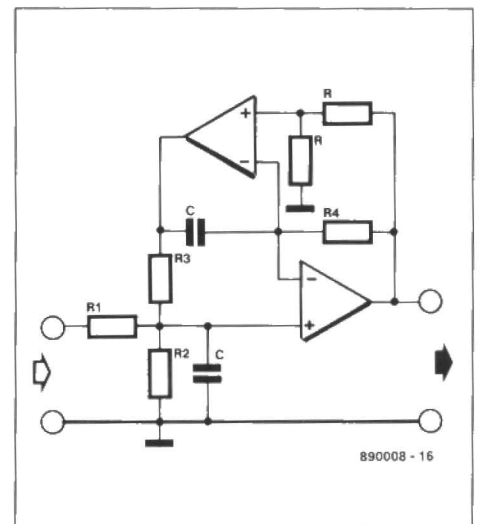
Fig. 31. This design, based on two opamps is suitable for filters that need a Q-factor between 25 and 100.

The maximum amplification of this configuration is $4Q_s^2$.

Next month we will start with a review of the various filter tables.

RADIO DATA SYSTEM (RDS) DEMODULATOR

sources: European Broadcasting Union; Philips Components
design: J. Bareford

After several years of study and practical tests, the Radio Data System (RDS) is currently implemented on all BBC/IBA VHF FM broadcast transmitters in the United Kingdom. Other countries in Europe are following now that all technical and functional specifications of the system have been finalized by the European Broadcasting Union (EBU). The primary aim of the RDS system is to permit the realization of automatic tuning features in car radios. This article provides an introduction to the operation of RDS, and presents a demodulator as a basis for experimental reception of this new service.

Most motorists are well aware that reception of VHF FM signals is troublesome in a moving car because of signal reflection, the limited range of the transmitters, and noise induced by the engine and other moving parts. The Radio Data System should alleviate some of the plight motorists suffer in trying to stay tuned to a particular programme during the trip. Especially in hilly countryside, and in areas with a high transmitter density, continued reception of a particular programme means frequent retuning of the car radio and a possible hazard for the motorist.

If so programmed, a RDS-compatible radio tuned to, say, BBC Radio-1, will produce this programme irrespective of the whereabouts of the car in the country. During the trip it is automatically tuned to the strongest signal available from a VHF FM transmitter that carries the BBC Radio-1 programme.

In addition to automatic tuning, RDS provides a host of other facilities which are of great significance in view of the increase in road traffic and the economic losses caused by traffic jams. Because they are linked by a microwave network, local radio stations such as Radio Kent can relay traffic information and emergency calls to 'RDS-compatible' motorists listening to a national service such as BBC Radio-1: when the traffic announcement is broadcast, the car radio is automatically tuned to the frequency used by the local station.
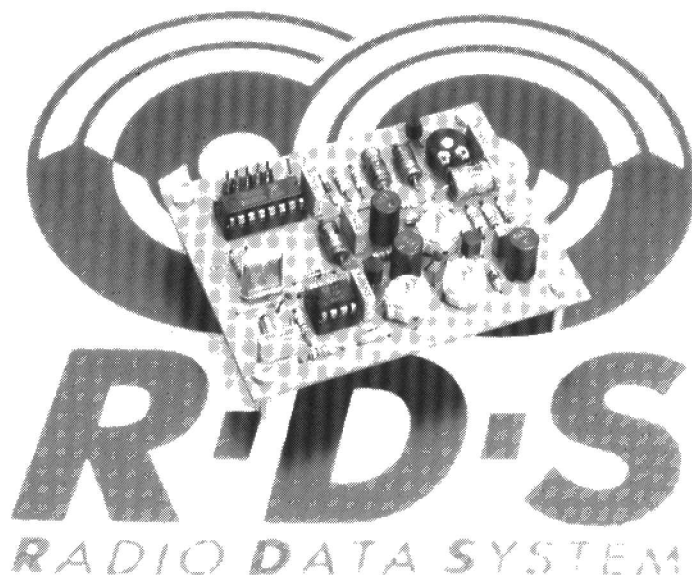
Features of RDS

The Radio Data System has been designed for use throughout Europe, and with future extensions in mind. This does not mean, however, that all technical features listed in Table 1 are available in a particular country.

Primary features

The Programme Service (PS) name and the Programme Identification (PI) are the most important of the *primary features* of RDS.

■ **PI** information enables the receiver to distinguish between countries or areas in which the same programme is transmitted, and the identification of the programme itself. The code, which is not intended to be displayed, enables the receiver to search for the strongest transmitter signal available for providing a particular radio programme in the car. The criteria for re-tuning the receiver would be the availability of a better signal having the same PI code.

■ **PS** information is a text consisting of a maximum of ten alphanumeric characters that form the station name or a meaningful abbreviation thereof, e.g. RADIO MERC for Radio Mercury.

Further important primary features include:

■ **AF** (alternative frequency list)
A list of up to 25 frequencies of VHF FM stations broadcasting the same PI code in one area or adjacent areas may be transmitted. The list can be loaded and stored in the RDS-compatible receiver.

■ **TP/TA** (Traffic Programme/Traffic Announcement)
These are essentially on/off codes that indicate whether a particular transmitter carries traffic information or not. The signal may be used for interrupting a cassette or compact-disc player, turning on a lamp or another signalling device in the car, or be taken into account during automatic frequency search.

Secondary features

A number of so-called *secondary features* are available for broadcasters. Again, not all of these may be actually implemented, although most are in the UK.

ON (Other Networks)
This code supplies information on programmes available on other networks. It may be used in conjunction with the AF list and the PI code to switch the receiver automatically from, say, BBC Radio-1 to a station broadcasting traffic announcements, say, Radio Kent.

CT (Clock Time and date)
UTC time codes are transmitted for conversion to local time in the receiver.

PTY (Programme Type)
This code identifies the programme type within 31 possibilities. Examples of PTY codes are: 00001 = News; 00101 = Education; 01110 = Folk Music. The PTY code may be used for search tuning.

PIN (Programme Item Number)
This code enables receivers and recorders to respond to preselected programmes or programme items. Use is made of the scheduled programme time.

RT (Radio Teletext)
Text transmissions are primarily intended for suitably equipped home receivers. In a car, however, they present a possible hazard, and are therefore intended for controlling a speech synthesizer.

TDC (Transparent Data Channel)
This oddly named code refers to the use of displays and computer equipment for downloading text or short computer programmes supplied by the RDS decoder.
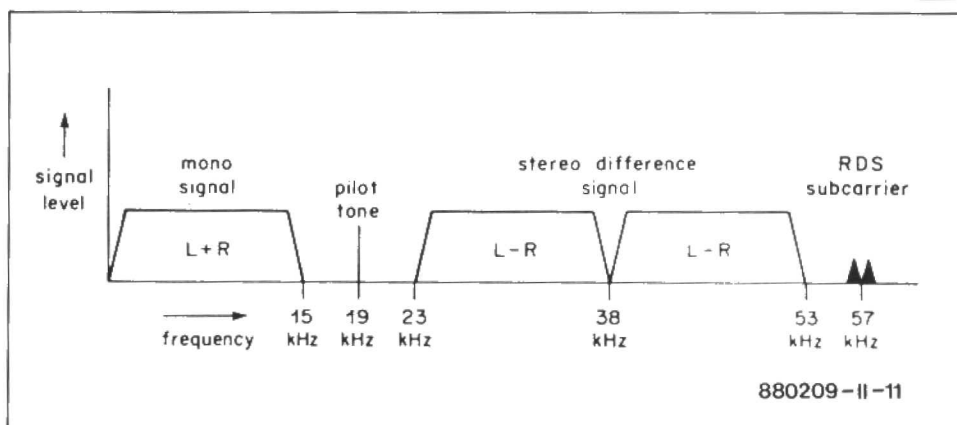
Fig. 1. Spectrum of the baseband transmitted by a VHF FM broadcast transmitter modulated with a conventional stereo multiplex signal plus associated pilot tone, and the RDS datastream.

DI (Decoder Identification)
These 4 coded bits indicate which of 16 possible audio processing modes (Dolby, surround-sound, HighCom, CX, and, possibly, de-scrambling) is applicable to the received programme.

MS (Music/Speech)
When used, this 4-bit code allows the receiver to distinguish between music and speech, and to control the volume accordingly.

IH (In-House)
This code is provided for use by the broadcasting authority only. Applications include identification of transmission origin, remote switching of networks or local transmitters, and paging of staff. Standard RDS receivers ignore IH codes.

Additional features:
In addition to the primary and secondary features discussed above, the relevant EBU recommendation lists two 'additional features':
Radio Paging, **RP**, may also be implemented as a separate feature of RDS, i.e. independent of IH codes. Similarly **TMC**, Traffic Message Channel codes, may be used independently of TP codes.

Technical description of the RDS signal
The RDS datastream is available continuously on a subcarrier added to the baseband transmitted by a VHF FM broadcast station. This baseband, which was formerly called the *multiplex signal*, already comprises the stereo difference signals $L-R$ and $L+R$ plus the associated pilot tone at 19 kHz (CCIR recommendation 450-1).
The RDS system uses a subcarrier frequency at 57 kHz, which is locked either in phase or in quadrature on to the third harmonic of the 19 kHz pilot tone. The resulting baseband spectrum is shown in Fig. 1.
The RDS subcarrier is suppressed and amplitude-modulated by a so-called *bi-*

phase coded data signal, whose structure will be reverted to below. The modulation system for RDS is therefore *double-sideband* with *suppressed carrier* (DSSC). Alternatively, it may be qualified as two-phase PSK (*phase shift keying*) with a phase deviation of 90°. Provision has been made to ensure that the RDS system does not interfere with the ARI (*Autofahrer Rundfunk Infor-*
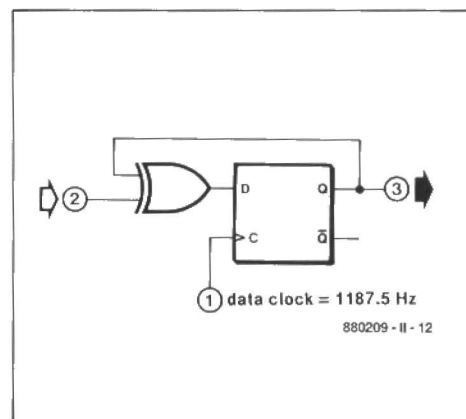
Fig. 2. Basic differential encoder made from a D-type bistable and a XOR gate.
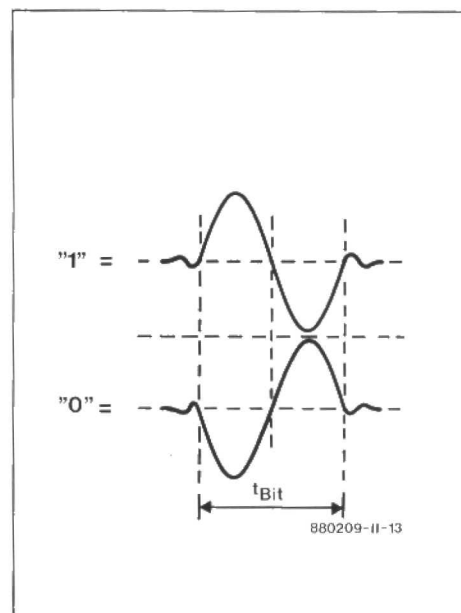
Fig. 3. Biphase-modulated symbols generated by logic 1s and 0s are composed of a positive and a negative half-cycle.

Table 1.
Features of RDS (EBU Recommendation)

Primary features	
PI	Programme identification
PS	Programme service name
AF	Alternative frequencies
TP/TA	Traffic programme/traffic announcement

Secondary features	
ON	Other networks
CT	Clock time and date
PTY	Programme type
PIN	Programme item number
RT	Radiotext
TDC	Transparent data channel
DI	Decoder identification
MS	Music/speech switch
IH	Inhouse information

Additional features	
RP	Radio paging
TMC	Traffic message channel
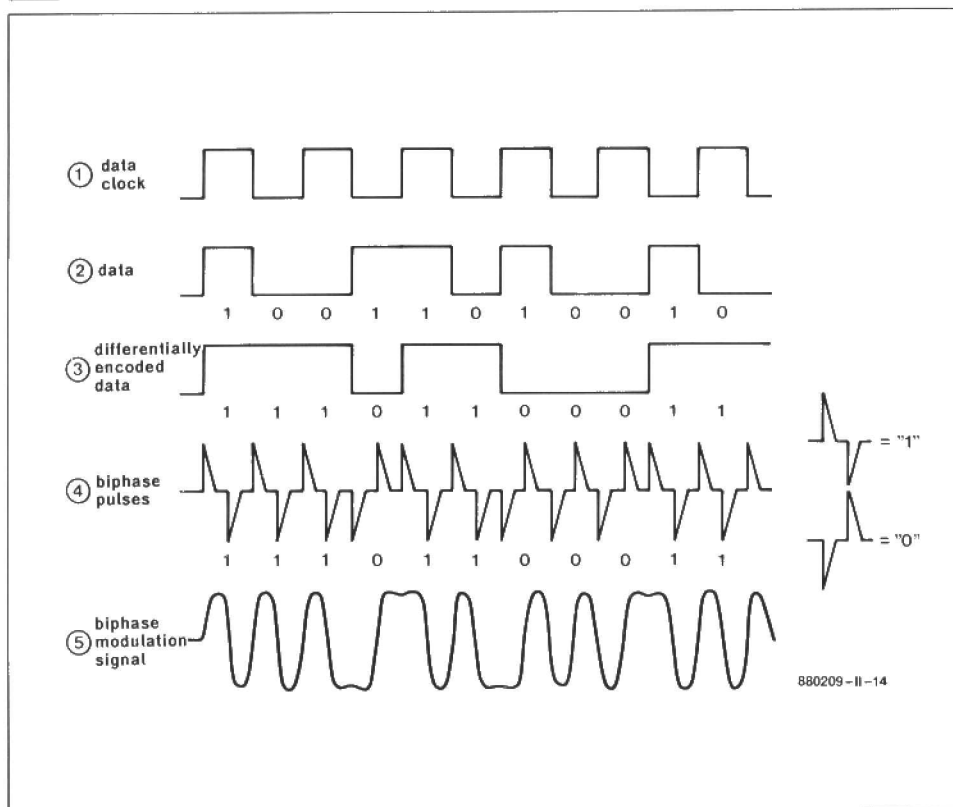
Fig. 4. Timing diagram illustrating the generation of the RDS datastream.

Fig. 4. In pure biphase coding, a logic 1 in the source data results in a positive half cycle followed by a negative half cycle (see Fig. 3). For a logic 0, the order is reversed. In practice, RDS uses a slightly different method based on biphase pulses instead of half-cycles (see the inset waveforms to the right of diagram 4 in Fig. 4). A logic 1 then corresponds to a positive pulse followed by a negative pulse. The opposite applies to a logic 0 in the source data.

After shaping and filtering to reduce the overall bandwidth, the biphase-coded RDS signal is modulated as shown in diagram 5 in Fig. 4.

In an RDS-compatible receiver, the bit clock and the NRZ data are recovered from the baseband by a special demodulator. In fact, filtering and shaping of the data spectrum is split equally between the transmitter and the receiver to ensure maximum noise immunity. Thanks to the differential coding, data is always decoded correctly irrespective of logic inversion in the receiver.

mation; Broadcast information for motorists) which has been in use for a number of years already in Federal Germany. The ARI system shares the 57 kHz subcarrier with the new RDS service.

The bandwidth occupied by the RDS sidebands in the baseband spectrum corresponds roughly to the rate of the databits, 1187.5/s. The data clock (signal 1 in timing diagram Fig. 4) is obtained by dividing the subcarrier frequency, 57 kHz, by 48.

Before it is applied to the RDS modulator, the datastream is coded differentially to reduce the overall bandwidth, to

minimize power waste by the subcarrier, and to ensure optimum reliability of the system. The basic layout of the differential encoder used is shown in Fig. 2. This circuit, based on a D-bistable with XOR logic feedback, is fed with the data clock signal and the NRZ (*non-return to zero*) data signal (diagram 2 in Fig. 4.). A NRZ signal remains logic high with subsequent logic high input pulses, and toggles only on the high-to-low or low-to-high transitions. The differentially coded signal is shown in diagram 3 in Fig. 4: the output state of the bistable changes only when the input bit is logic 1. The differential logic signal is then biphase-coded as shown in diagram 4 in

Coding structure of the RDS datastream

RDS data is subdivided in 16-bit words, accompanied by a 10-bit checkword. The MSB (*most significant bit*) is always transmitted first. The data and checkword bits together form a 26-bit *block* (Fig. 5a). Four of these blocks form a 104-bit *group*. Since all data is transmitted continuously without gaps between the blocks or groups, the group transmission time is 87.5 ms.

The 10-bit checkword enables the receiver to detect and correct errors which may occur when reception is impaired or marginal. This checkword, composed of bits c'_n (Fig. 5) is of the so-called CRC (*cyclic redundancy check*) type. The 10-bit CRC word is obtained by multiplication of the 16-bit information word with a matrix polynomial, and subsequent XOR-ing with an off-set word. After these operations, the checksum is appended to the information word. The checkword is transmitted MSB-first. There are 5 off-set values, A, B, C, C' and D, appended to blocks 1, 2, 3, 3 and 4 respectively. These off-set values enable the receiver to synchronize with the received datastream and thus to process the groups and blocks separately.

Group types

Each group contains data that marks its status. This type-marking is achieved with the first 5 bits (MSBs 15 through 11) in block 2 of the group. The first four bits, 15 through 12, indicate the group number, the fifth bit the type, A or B. The group types so obtained carry the features listed in Table 1. All group types comprise the complete 16-bit
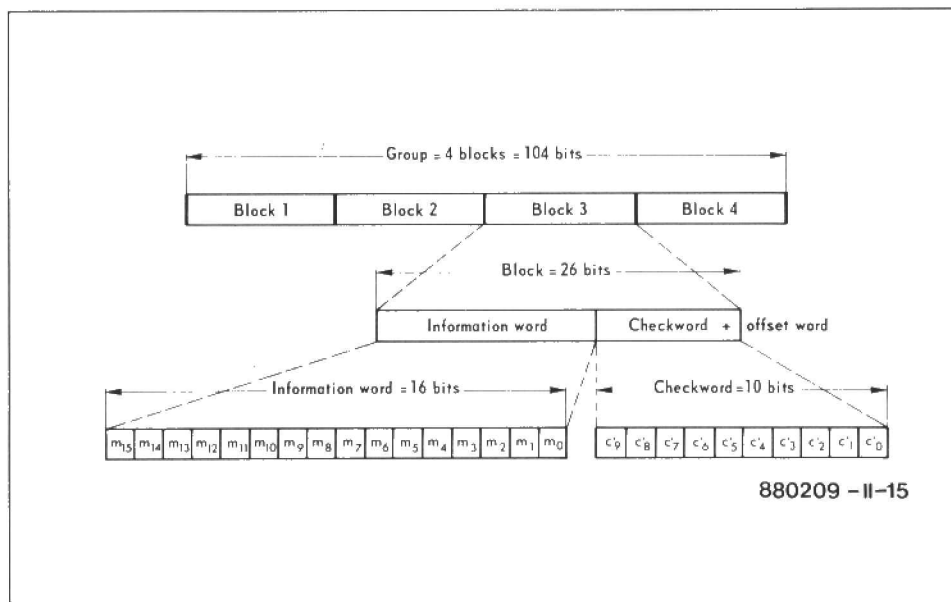
Fig. 5. Structural analysis of the RDS data format (*courtesy European Broadcasting Union, Brussels*).

Program Identification (PI) code in the first block, which thus has the highest repeat rate, and is received at 87.5-ms intervals. Similarly, the PTY (Programme Type) code and TP (Traffic Programme Identification) bits are assigned fixed positions in the second block, whose bit-order is basically as follows. Starting with the MSB (bit 15) the bit-order is:

■ the 4-bit group number (0 to 15);
■ group type (1 bit) A or B;;
■ TP flag (1 bit);
■ PTY (5 bits);
■ M/S flag (1 bit);
■ DI (1 bit);
■ decoder address (2 bits).

Alternatively, the last 3 bits indicate which of the eight PS characters are contained in the 2 bytes of the fourth block in the group. In the B-version of all groups, blocks 1 and 3 contain PI data. This state is marked with off-set word C' instead of C. In the A-version of group 0, the third block contains sixteen AF bits.

Although bound by EBU recommendations, broadcasting authorities are in principle free to decide on the order and repetition rate of the groups. The use of the bit positions reserved for PTY and M/S is optional, and dummy bits may be transmitted when these services are not used.

Towards a decoder

Apart from RF and AF circuits, an RDS-compatible radio requires the following modules:

■ an RDS demodulator that recovers the NRZ data and the bit clock from the baseband;
■ a decoder capable of (1) recognizing the block structure by synchronizing with the off-set masks, and (2) error correction;
■ an alphanumerical display unit;
■ an interface for controlling RF and AF sub-circuits in the radio.

Fig. 6. The RDS concept proposed by Siemens is based on two chips for demodulation (SDA1000) and processing (SDA1010) of the composite RDS signal. The serial output of the SDA1010 supplies a data signal suitable for feeding to the system processor in the (car) radio.

The first module, the demodulator, is described further on in this article. For evident reasons, this demodulator takes its input signal from a point ahead of the de-emphasis circuit.
A number of semiconductor manufacturers, including Siemens and Valvo (Philips Components West-Germany), have recently released integrated RDS demodulators. The use of one of these, the SAA7579T from Valvo, is described in an application intended for further experiments with RDS. Meanwhile, Siemens have announced a two-chip set, the SDA1000-SDA1010 for RDS demodulation and partial decoding (see Fig. 6). The final link, an intelligent, that is, programmable and error-correcting, display driver with radio interface, however, does not seem to be available from any manufacturer as yet, leaving ample room for experiments and developments with software in general and single-chip microcontrollers in particular.

For your experiments: an RDS demodulator

The Type SAA7579T is a CMOS RDS decoder in surface-mount technology. The chip is manufactured by Philips Components. The circuit proposed here differs from the standard application suggested by the manufacturer. For various reasons, this application was found to give unsatisfactory results, so that a better configuration was developed.
The function of the RDS demodulator is the recovery of the clock and data signals from the encoded RDS bitstream. As already discussed, the serial RDS signal is first coded differentially and then converted to biphase symbols that are modulated on to a 57 kHz subcarrier. The suppressed RDS carrier and the sidebands caused by the

modulation process are broadcast by a VHF FM transmitter, together with the stereo multiplex signal.
In a receiver, the multiplex signal is available at the output of the FM demodulator. From there, it is fed to the stereo decoder and the parallel-connected RDS decoder. The RDS decoder has a bandfilter that extracts the RDS sidebands at 57 kHz from the baseband. A synchronous demodulator is then used to recover the biphase RDS signal and the subcarrier. Next, the biphase symbols are decoded and restored to their original bit format with the aid of a differential decoder.
This brief description covers the basic operation of the hardware required for demodulating and decoding the RDS signal. The remaining functions of word-synchronization, group and block recognition on the basis of off-set word evaluation, error correction, data processing and interfacing to the radio and display require the use of a microprocessor and, of course, a suitable control program.

SAA7579T: functional description

That the SAA7579T comes in a surface-mount assembly (SMA) package is not surprising because the chip is intended for applications in car radios. The functional structure of the chip and its pinning are shown in Figs. 7 and 8 respectively.
The SAA7579T works from a single supply voltage of 5 V, and draws only 1.6 mA (typ.), The pre-filtered RDS signal is applied to input pin 15, designated MUX. A so-called Costas-loop with associated 4.332 MHz crystal oscillator demodulates the signal synchronously, and also supplies the regenerated 57 kHz subcarrier. The presence of an ARI signal is flagged by a high level at pin 1.

Table 2. RDS Group Types

group type	information bits
0A	PI, PTY, TP, TA, DI, MS, PS, AF
0B	PI, PTY, TP, TA, DI, MS, PS
1A/B	PI, PTY, TP, PIN
2A/B	PI, PTY, TP, RT
3A/B	PI, PTY, TP, ON
4A	PI, PTY, TP, CT
5A/B	PI, PTY, TP, TDC
6A/B	PI, PTY, TP, INH
15B	PI, PTY, TP, TA, DI, MS
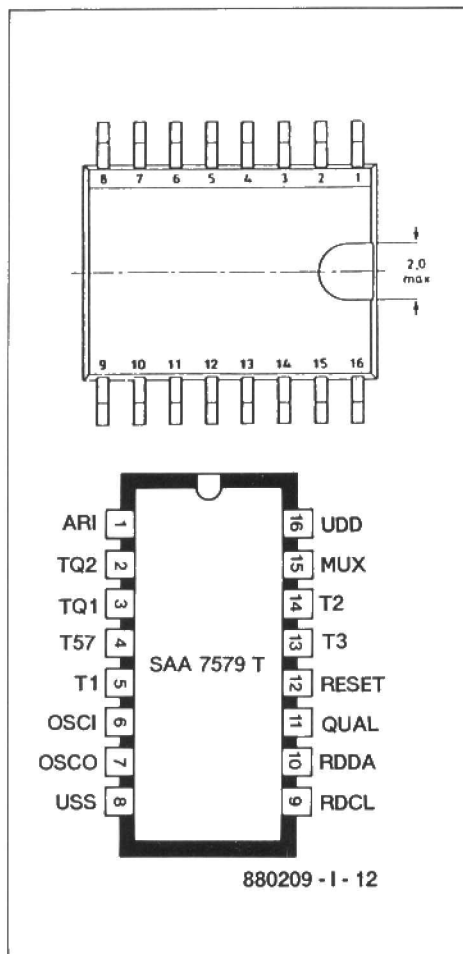
Fig. 8. Pinning of the SAA7579T RDS demodulator in a 16-way surface-mount enclosure.

The Costas-loop supplies the regenerated 57 kHz subcarrier and the demodulated RDS signal, which is a synchronous stream of biphase-coded symbols, at a rate of 1187.5 Hz. The regenerated subcarrier is divided by 48 to supply the data clock signal, RDCL, which has a frequency of 1187.5 Hz. Synchronization to the data in the biphase bitstream is achieved by controlling a programmable divider. The process of data recovery is continued in a biphase symbol decoder and a subsequent differential-phase decoder.

The use of differential encoding makes the demodulation process independent of the absolute phase of the modulation signal, which results in a high degree of immunity of the system to inversion of the signal in the receiver/demodulator.

The RDS datastream is synchronized with the clock signal and then fed out to pin 10, marked RDDA. The signal is digitally compatible to simplify direct interfacing to a microprocessor system. The same applies to the auxiliary signals, the clock, RDCL (pin 9), and the quality bit, QUAL (pin 11). The RDS data should not be processed when QUAL is logic low.

Relative to the clock, data are valid for 417 μs after a clock transition, irrespective whether referenced against the positive or negative pulse transition.

Table 3. Pinning of SAA 7579T

Pin	Signal	Function
(1)	ARI	ARI indication output (ARI = H)
(2)	TQ2	Test output 2
(3)	TQ1	Test output 1
(4)	T57	57 kHz test output
(5)	T1	Test input 1
(6)	OSCI	Oscillator input
(7)	OSCO	Oscillator output
(8)	Uss	Ground (0 V)
(9)	RDCL	RDS clock output
(10)	RDDA	RDS data output
(11)	QUAL	Signal quality indication (H = o.k.)
(12)	RESET	Reset input for test logic
(13)	T3	Test input 3
(14)	T2	Test input 2
(15)	MUX	RDS input (MUX; clamped and filtered)
(16)	UDD	+5 V supply voltage

This forces data changes, i.e., toggling of the data signal (RDDA), to take place 4 μs before a clock transition (positive or negative).

Application circuits

The application circuit for the SAA7579T developed by Philips Components is given in Fig. 9. Transistor T_1 forms an amplifier with relatively high input impedance. The high-Z input amplifier ensures that the FM demodulator in the receiver is not excessively loaded. The amplified baseband is taken through a 4-stage bandfilter with inductors from Toko. The frequency response and group delay characteristic of this filter are shown in Fig. 10. The RDS signal at the output of the filter is applied to a Type TBA120U which functions as a limiter/amplifier. Buffer T_2 raises the composite RDS signal to a digital level so that it can be applied to input MUX of the SAA7579T.

The above application circuit for the RDS decoder was tested and subsequently modified. The final version developed in the *Elektor Electronics* laboratory is shown in Fig. 11.

The amplification of the first stage is adjustable with preset P_1. Good selectivity at 57 kHz is achieved with a three-stage bandfilter based on fixed inductors with parallel, partly adjustable, capacitances. Potential divider R_6-R_7 provides the DC bias for emitter follower T_2. This stage ensures that the bandfilter is only lightly loaded while driving the input of fast opamp IC_1 with sufficient signal amplitude. Diode D_1 compensates temperature drift of the transistor's U_{B-E} parameter. The opamp, a Type LF357, is set up to provide an amplification of 101. Series network R_{11}-D_2 forms a clamping circuit to limit the output signal of the opamp to digital levels (approx. 0 V/+5.5 V). The component configuration around the RDS decoder chip
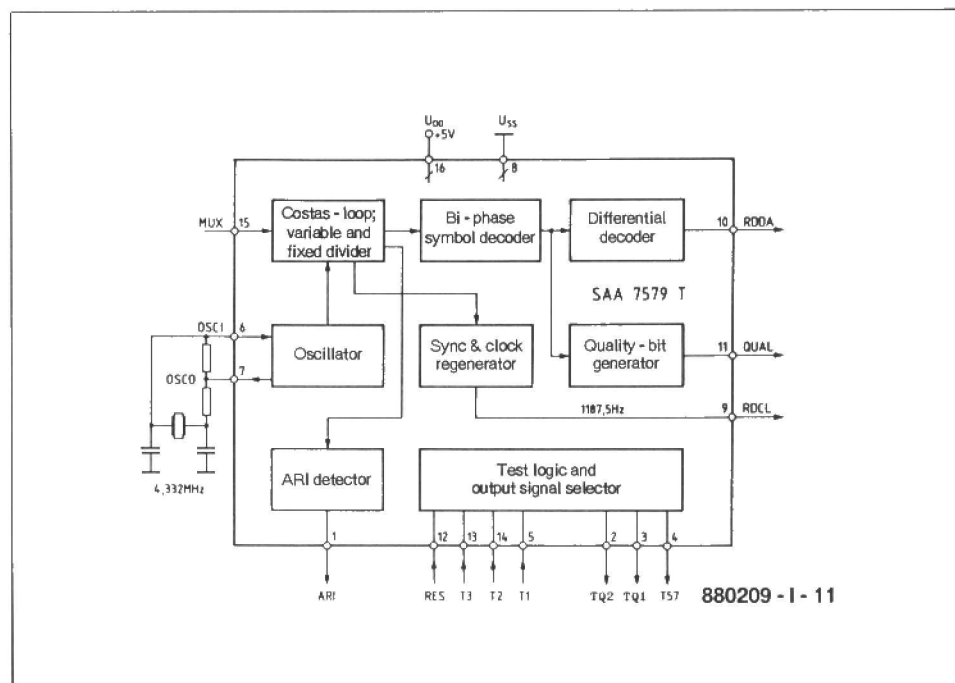
Fig. 7. Block schematic of the SAA7579T (*courtesy Philips Components*).
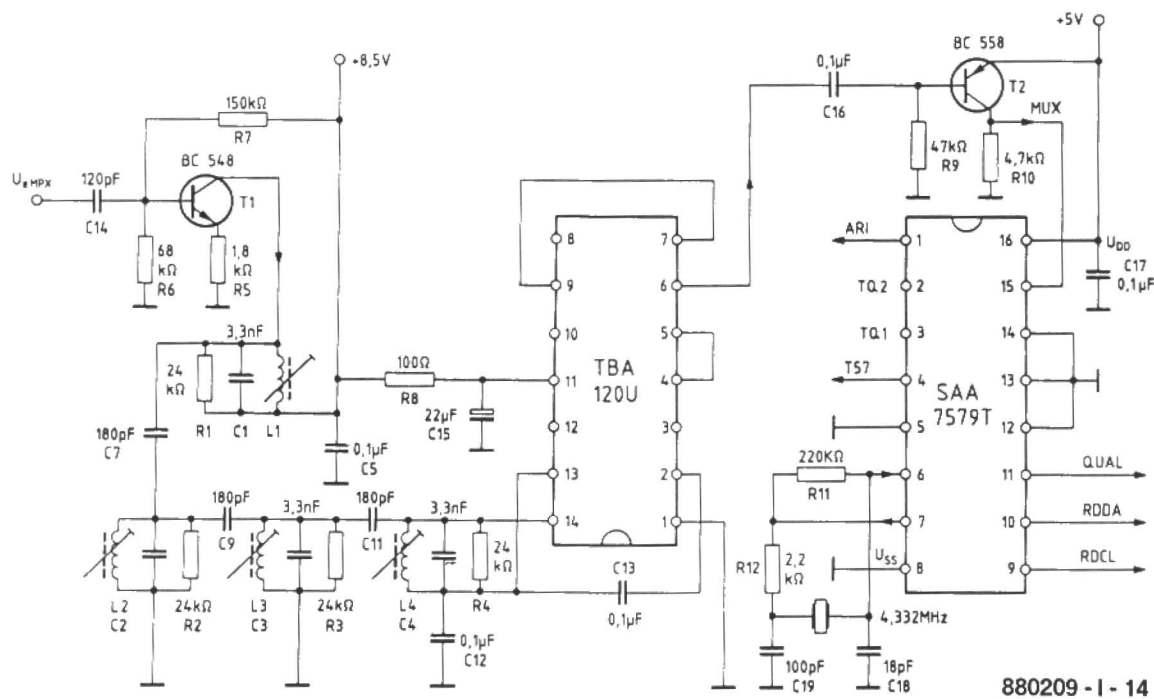
Fig. 9. Original application circuit for the SAA7579T. The Toko inductors in the four-stage band-pass filter tuned to 57 kHz are Type 126ANS/A3561HM.

is standard and identical to the original Philips Components application. All outputs of the SAA7579T, and the regenerated 57 kHz signal, are buffered with a CD4050 to enable driving a microprocessor port.

The RDS decoder chip and the input stage around T1 are fed from a regulated +5 V rail provided by on-board regulator IC3. The opamp must be powered from a higher, but also regulated, voltage, in this case +12 V available at the input of the regulator. The total current drain of the circuit is lower than 100 mA.

Alignment and application

Little needs to be said about the alignment of the demodulator. The amplification of the input stage is adjusted with P1 until an oscilloscope shows that D2 starts to clip the signal. This happens

Fig. 11. Circuit diagram of the RDS demodulator.
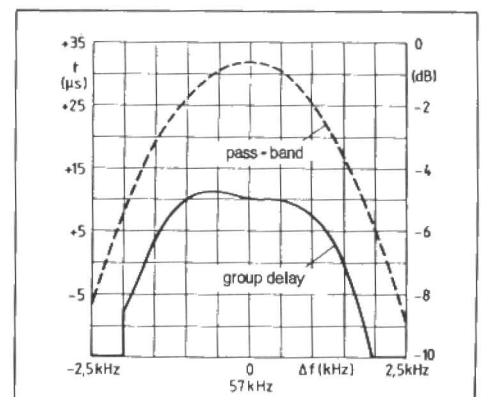
Fig. 10. Pass-band characteristic and group delay plots of the 4-stage band-pass filter.

Parts list

Resistors (± 5%):
R1;R13 = 220K
R2;R10 = 100K
R3 = 6K8
R4;R8 = 4K7
R5;R9 = 1K0
R6;R7;R11;R12 = 2K2
P1 = 100K preset H

Capacitors:
C1;C12 = 10n
C2;C7;C10 = 40p or 45p trimmer capacitor
C3;C8;C11 = 47p; polystyrene
C4 = 47µ; 16 V
C5;C13;C19 = 100n
C6 = 2p7
C9 = 4n7
C14;C17;C18 = 1µ0; 16 V
C15 = 100p; polystyrene
C16 = 18p; polystyrene

Inductors:
L1;L2;L3 = 68 mH radial inductor with ferrite encapsulation

Semiconductors:
D1;D2 = 1N4148
T1 = BF451
T2 = BF494
IC1 = LF357
IC2 = SAA7579T (Philips Components)
IC3 = 78L05
IC4 = 4050

Miscellaneous:
X1 = quartz crystal 4.332 MHz; parallel resonance.
PCB 880209 (not available through the Readers Services).

Fig. 12. Track layout and component overlay of the printed-circuit board designed for the experimental RDS demodulator.

when the opamp supplies a signal with a swing greater than 10 V. The trimmers in the bandfilter are adjusted for maximum amplitude of the signal at the emitter of T2. The QUAL bit indicates whether valid data are being received or not. The ARI bit will not go high in the UK since there are no transmitters broadcasting this service.

The signals are then ready to be applied to a microprocessor-system running an appropriate program. How about designing such a system as your final-year course project? Let us know!

For further reading:

1. EBU Technical Document 3244-E: *Specifications of the radio data system RDS for VHF/FM sound broadcasting.* European Broadcasting Union, Brussels, March 1984.

2. EBU Technical Document 3260: *Guidelines for the implementation of the RDS-system.* European Broadcasting Union, Brussels, January 1988.

3. Parnall, S.J., Lewis, A.R., and Robinson, J. (BBC): *The technical realization of a new broadcast service.* Paper delivered on the occasion of IBC 1988, Brighton.

A CLOSER LOOK AT THE TRANSPUTER

by Pete Chown

The transputer is probably the most powerful microprocessor available today; ironically, it is also one of the least used. It is a RISC chip and a very powerful one. Its power stems mainly from its ability to join transputers and use them as a single unit, thus multiplying the power. The fastest transputers available today have a speed of 15 MIPs (millions of instructions per second) and 2–4 Mflops (millions of floating points operations).

If you built a computer with a transputer, it would be about the fastest desktop computer you could buy. It would be several times faster than a VAX 11/780. if you used four of these devices, your (still desktop) computer would be faster than some mainframes with a speed equal to about 1/10 of a Cray-I. If you used fifty of the devices you would reach the bottom of the supercomputer range (remember that these could still fit under a desk). A thousand of the devices would form the fastest computer ever built, at a cost around that of a small mainframe.

Remember the Mandelbrot plots that took 3 hours on a BBC Micro? One transputer would do this in 20 seconds. Fifty transputers would do it fast enough to zoom in and out and move in real time!

Transputer architecture
If this all sounds too good to be true, there is a catch. The transputer is very difficult to program, because you must tell it explicitly which parts of your program may be run at the same time on different transputers. This inevitably leads to the other problem: that programs written on one computer can be hard to make run on a different one, because there are different numbers of transputers available. To get round this problem, the transputer has a feature whereby more than one concurrent process may be run on one transputer. This means that you can split your program into a lot of processes. Then if there are a lot of transputers available, the processes are run on separate ones, otherwise on the same one. You are, however, still left with that extra level of complexity in having to decide which parts of your program may be run at the

same time.

The transputer is a complete computer on a chip: it incorporates 2 or 4 K of memory, microprocessor and interfacing and it is here that the next problem arises. If you add external memory, as you obviously need to do, the speed will be about halved. This still, of course, leaves it faster than most other microprocessors. You can also add additional transputers if maintaining the speed is vital.
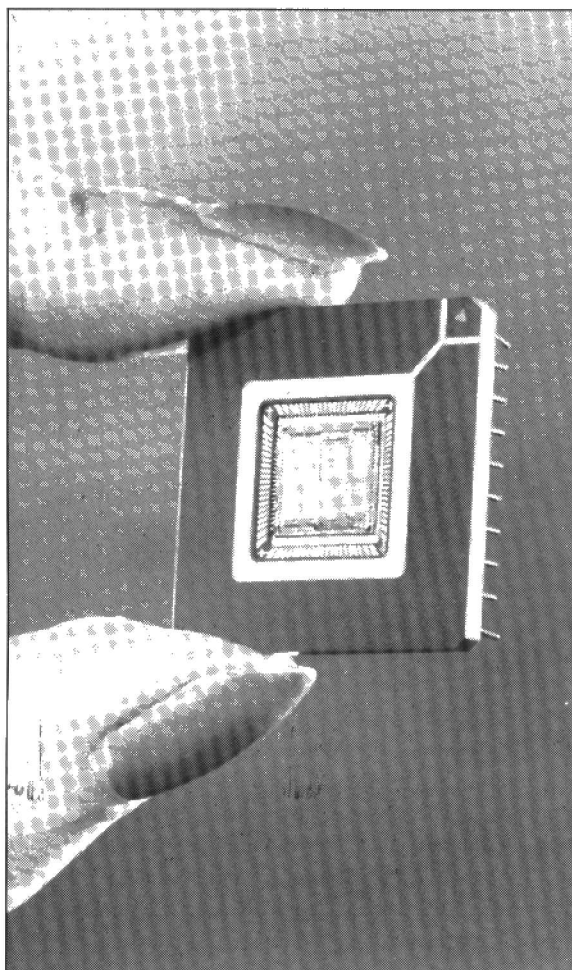
Inmos (who make the transputer; not to be confused with Intel, who make the 8086 series processors) have provided a language for programming it, called Occam, which has the facilities for parallel processing (that is, running processes at the same time on different transputers). Unfortunately, it is awful. To give you some idea, the level of indentation at the left margin is significant in determining when statement blocks end. To fill the gap, therefore, conventional languages have been produced with parallel constructs. How good these are remains to be seen: Occam forces you at every stage to think about possible parallel processing, but the others, such as parallel–C, may not and may allow you to avoid really having any parallelism to speak of, not by choice but just by not thinking.

The transputer, like any microprocessor family, comes in a variety of speeds and bus widths. Basically, the T212 series has a 16-bit bus width, the T414 a 32-bit one, while the T800 is the same as the T414, but with a floating-point processor built in. Table 1 lists all the processors with their speeds that are currently made.

The Inmos link system
Obviously, with many transputers working on a problem at the same time, communications is needed. Equally obviously, it can not be by a conventional network, because the more transputers that are added, the longer the delays will become.

Inmos therefore decided to have serial links that join either two transputers or a transputer and an I/O device. These links

are bi-directional and operate at 5 or 10 MHz; 5 MHz is the standard, but many of the devices have an option to run faster. There are 4 links provided from each transputer and one on each I/O device.

Interfacing to I/O devices is much simpler with the links than with conventional memory-mapped I/O. Unfortunately, four links are seldom enough and so a memory-mapped system must be used as well. There are specialized interface chips for the transputers as with any microprocessor family and these allow links to be interfaced to a bus on another microprocessor, or to control a printer port. Some of the support devices can be programmed in their own right and are used for controlling discs or other devices needing complex controllers.

Transputer networks
Networks of transputers may be built up, so that a message can travel from one transputer to another; although there may not be a direct route, it gets there through several other transputers. It might seem that designing an efficient network would be difficult, but in fact the difference in performance between the best and the worst is usually only about 10%. In very large systems, it becomes more important, but as long as a reasonable configuration is chosen, there is unlikely to be a problem. Some common arrangements are shown in Fig. 1 and Fig. 2.

Programming in parallel
Occam has been designed from the outset with parallel processing in mind. In each statement block, you must decide whether the statements are to be executed one after the other, at the same time, or whether the first statement to become possible is to be executed. The last sounds somewhat useless, but is, in fact, used when, for example, a program is displaying a menu and at the same time it is spooling to a printer. Here it is possible to say that if a key is pressed first, it will be acted on, whereas if the printer buffer becomes empty, new text will be supplied. The program in this example would then loop back to deal with the next key press or requirement for text.

Four other languages exist for parallel processing: assembler; parallel-Pascal; parallel-C; and parallel-Fortran. Assembler on the transputer is very complex and its use can not be recommended. There is little point in using Fortran since it is rather out-

Processor	Bus	MIPs	Floating point
IMS T212A-G17S	16-bit	8.75	no
IMS T212A-G20S	16-bit	10.00	no
IMS T414B-G15S	32-bit	7.50	no
IMS T414B-G20S	32-bit	10.00	no
IMS T800C-G17S	32-bit	8.75	yes
IMS T800C-G20S	32-bit	10.00	yes
IMS T800C-G30S	32-bit	15.00	yes

Table 1. Transputers and selected characteristics currently in production.

dated. Parallel-Pascal and parallel-C are the best, but it is dubious whether either of them is worth using for reasons already stated: programs will tend not to incorporate as much parallelism as they otherwise would.

Interfacing the transputer
Interfacing should be carried out if possible through Inmos links. If there are not enough links, memory-mapped I/O could be used, but Inmos standard devices would not work and devices with sufficient speed from other manufacturers are hard to find.

The basic interface device is the IMS C011A. This provides a means of interfacing it to bus systems and also controls the

Fig. 1. A node of transputers that leaves four links free and is useful for small systems. It can also increase the size of a network if each transputer is replaced by a n node of them.

Fig. 2. A hypercubic network, which is used where there are rather more transputers. It does not, however, leave any spare links, which may be a problem.

more basic I/O ports. The IMSM212B provides a more sophisticated interface for discs and other peripherals that normally need dedicated controllers. This is programmed by downloading a program into its link on power-up. It then uses a fairly standard transputer instruction set.

External memory will be needed in all but the simplest applications (an arithmetic processor for your micro might not, and it would be a lot faster than any other on the market!). External memory is added by a simple bus structure. This is multiplexed in almost the same way as the 8086 bus, except that it has 32-bit data and addressing.

Many of the current transputer systems use an interface to a bus system, more specifically to an 8086-type bus. The development system supplied by Inmos plugs into the expansion bus on an IMS-compatible PC. More recently, another system has appeared, which is the fastest desktop computer ever built. It is also an IBM-compatible that can run PC-DOS and OS/2 programs. This is not its main application, however, and it relegates its 80836 processor to an unaccustomed back seat!

It is interesting to note how the clocks are provided. Distributing high-frequency clocks produces problems of crosstalk and r.f. interference. Consequently, a single 5 MHz clock is sent to all transputers, regardless of speed, and an internal resonator steps up the clock speed to the 17–30 MHz required for the internal operation of the parts.

Transputers do not have an interrupt system like a more conventional microprocessor. Instead, they have an event line. A single process inside the transputer can be triggered by the event line at any one time. When the process has started, the transputer offers a high output on its event acknowledge line. This then allows the process to determine the cause of the event and to act on it.

Starting a transputer on power-up is different from doing it with other microprocessors. You have a choice of two ways of doing it: you can have the BootFromROM line low, in which case the transputer will download a program from a link and start up, or you can have BootFromROM high, in which case the transputer will commence executing two bytes before the end of memory. This would then contain a back-
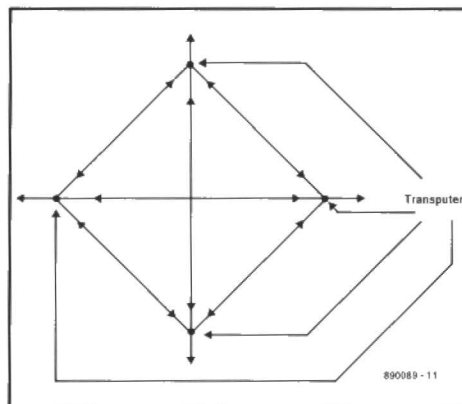
ward jump to the start of the program in ROM.
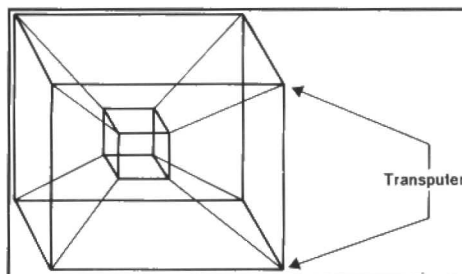
Occam

The language Occam is the closest Inmos recommend that you should get to assembly language. Generally, Occam statements compile almost exactly into machine code instructions, but this is not always the case. One notable exception is the floating-point operations on transputers without a built-in floating-point processor. There is one exception to Inmos's recommendation that you should not program in assembly language and this is for compilers: they do not advocate languages compiling into Occam.

The transputer is so completely different in its parallelism from what has gone before that it is necessary to develop completely new languages to deal with it. Ultimately, languages such as Prolog may be used, but Prolog has not had long enough to evolve to become really usable, except for some particular types of problem.

ALT, SEQ, and PAR

ALT, SEQ, and PAR form the Occam equivalent of Pascal's BEGIN and start a statement block. Statements following them that are in the statement block must be indented and the block is concluded by having a statement that is outside the indentation.

SEQ is the most easily understood Occam block statement. It tells the compiler that the statements in the block are to be executed in sequential order. This would seem obvious as far as a conventional computer is concerned, but must be specified with transputers.

PAR is the equivalent of SEQ but tells the compiler to excecute all the statements in the block at the same time if this would help to improve efficiency. They will not necessarily be executed at the same time, because if there are not enough transputers to allocate one per statement, one transputer may swap between them as concurrent processes.

ALT is a statement peculiar to Occam that is not really parallel in nature. It has some analogue in the asynchronous traps that can be used on a VAX to restart a process on a particular event occurring, or perhaps on the interrupts that can be trapped on micros, but it goes far beyond either of these. What it does is to suspend execution of the current process until one of the statements in the ALT construct can execute

(they will begin by being unable to execute because they will all be waiting for some kind of input, a buffer to empty, the event line to be asserted, or some other similar occurrence). When one can execute, it runs, but all the others are abandoned. It is thus possible to have a process running, but not consuming any processor time unless some input occurs that must be dealt with by it. In the mean time, foreground tasks may continue as normal.

Channels

An Occam channel is the means of communication between processes running on the same transputer, or processes running on a separate but linked transputer. They do not handle routeing across a network by themselves: this would be handled by the operating system in a complete transputer system.

Channels are declared by the statement:

CHAN OF (type) (channel name):
(process)

They are then valid for the process immediately following the declaration. To use a channel, one process created by the process following the declaration (with a PAR statement) would write into the channel. It would then be suspended until the value had been read out of the channel again by another process. It could then carry on and possibly write another value in. Conversely, a process might attempt to read the channel first, and in this instance it would be suspended until another process had written a value in it, and it could then proceed.

Channels can also work across transputers. In this case, the channel declaration is made to refer to the link instead of to a word in memory. This is not difficult because the links are memory-mapped within the transputer chip. Generally, this would not be done until run-time: in this case, the same program could run on a micro or a supercomputer, and all that would be changed would be the number of processes running on any one transputer.

It is possible to ensure that certain processes run on different transputers, for example to guarantee that a minimum level of performance is provided: there is no point in running a simulation of the aerodynamics of a rocket (the type of problem for which the Cray-I was developed) on a computer consisting of a single transputer. If the outermost PAR construct in a program is replaced by PLACED PAR, each of

the component processes will be run on separate transputers. It will then be an error if there are more processes than transputers.

It must be emphasized that ways of allocating processes to transputers, and communicating between processes and transputers, are very much in the formative stage and, although the hardware is present, there are as yet no established software techniques for it. This is one of the great problems with transputers today, and indeed with any parallel processing machines, but no solutions seem to be forthcoming. Any solution would be infinitely preferable to no solution, but the emphasis appears to be on finding a perfect solution, rather than simply finding an answer that works, which is what everyone wants!

Variables, loops and procedures

It seems almost surprising that Occam has anything as mundane as variables, loops and procedures, but it does, although the loops are like nothing seen before.

The variables are declared for each process: they can then be accessed by any sub-process. The declarations take the form:

(type) (variable name): (process)

The permitted types are:

BYTE	numbers 0 to 255
INT/INT32	numbers –80000000 hex to +7FFFFFFF hex
INT16	numbers –8000 hex to +7FFF hex
INT64	numbers –8000000000000000 hex to +7FFFFFFFFFFFFFFF hex
REAL32	32-bit IEE real (8-bit exponent)
REAL64	64-bit IEE real (11-bit exponent)
BOOL	Boolean
TIMER	refers to the real-time clock on any transputer

An example of a variable declaration would therefore be:

INT X:
SEQ
Instructions referring to variable X
Instructions NOT referring to variable X.

Loops in Occam are arranged by creating a large array of parallel processes for a

loop where each iteration is performed at the same time, or by creating one process at a time for loops where the iterations must be done one after another. Examples of both these loops are easy to find. Suppose you were trying to find a square root iteratively: you would then need to use your result from the first time around the loop in the second. You would thus use a sequential loop to ensure that the right intermediate result was used. If, however, you were plotting a large number of points on the screen, it would make sense to do them at the same time. It is worth noting that sequential loops are slightly faster if there is nothing else to choose between them, for example, if there was a choice between a sequential loop and a parallel loop where the intermediate results were passed on via channels. Note that WHILE loops are always sequential for obvious reasons.

Loops are started as follows:

SEQ i = 0 FOR n
           sequential FOR type loop
(statements in loop)
(statements NOT in loop)

PAR i = 0 FOR n
           parallel FOR type loop
(statements in loop)
(statements NOT in loop)

WHILE (condition)
           sequential WHILE loop
(statements in loop)
(statements NOT in loop)

Before we consider procedures, it is worth considering how to avoid procedures. Remember that a procedure called only once is basically inefficient and would be included only to make the program more readable. To get around this, the Occam editor provided with the transputer development system incorporates a system of folds. This means that the listing you edit can be folded so that you only see the present level that you are working on. Levels below that (i.e., lower down in a top-down approach) are replaced by explanatory text. If you wish to see them for debugging, you press a key and move down a level. It may well be that Occam programs tend to be less procedural than other languages.

There will, of course, be times when the use of a procedure is desirable and Occam provides procedures for these situations. They are declared by something like:

```
PROC square (INT x)
    x: = x*x
:
```

The name is used to call the procedure, for example:

```
square (n)
```

Functions are also provided in Occam. The following factorial calculation will give some idea of the construction of Occam as well as of the use of functions.

```
INT FUNCTION factorial (INT n)
    INT p
    VALOF
    SEQ
        p : = 1
        SEQ i = 1 FOR n
        p : = p* i
    RESULT p
:
```

Note that a sequential loop was used, not because it would be impossible to use a parallel loop, but because the extra over-heads of a parallel loop were not justified by the small amount of processing being carried out inside it.

Notice that a colon follows the function and procedure declarations in the same way as it does variable declarations. This is because procedures and functions are regarded as being merely certain variable types.

Introducing machine code
As mentioned earlier, Inmos do not recommend the use of machine code other than for writing compilers. Many people have disregarded their advice, and a macro assembler is now available for the transputer development system. Transputer assembly language is no different from any other: programs will run more quickly, but development time will be far longer and programs may bog down more easily and become impossible to debug. Unfortunately, this happens more easily on the transputer. Suppose a location in memory is being corrupted. This type of bug may be nearly impossible to find in a conventional microprocessor. If hundreds of processes are going on at the same time when the corruption occurs, finding the correct one will be a hopeless task.

Actual machine code programming is beyond the scope of this article, but it is possible to give some idea of the way the transputer's instruction set works. Basically, the transputer uses reverse

polish notation, in the same way as Forth. This means that all operations take place on a stack, and there are instructions to load the stack and save results.

There are three registers in the transputer that form its arithmetic stack. There is no stack pointer. This means that the stack may never exceed three words, and it must never be allowed to under- or over-flow, since undefined effects will then occur: it is not detected as an error.

The transputer also has a workspace pointer that points to the local variables. It allows access to all the variables in a typical system because it would be arranged so that the variables declared at the lowest level immediately followed the workspace pointer, followed by the next lowest level of variables, and so on up to the globals furthest from the pointer.

There are also PC and operand registers. The operand register is used to construct literals and input addresses. This is because all transputer instructions are single byte. Consequently, literals will take several instructions: most literals, however, are small and positive or small and negative. The transputer can produce these with a minimum of instructions. Instructions that expect an operand have space for a parameter of 0–15 to be specified. The most common is LDC, which pushes a value on to the arithmetic stack. Some examples are the best way of understanding how the literals are coded (the # sign means hex):

LDC #02: LDC can take a single nibble as a parameter, so this can assemble directly to #42

LDC #12: this becomes PFIX #1 LDC #2, generating it in two stages; the code is #21; #42

LDC #–12: this is negative, so it becomes NFIX #1 LDC #2; the code is #61; #42

If larger positive or negative numbers are required, additional prefix instructions are added. For negative numbers, only one NFIX is required and the rest can be PFIX instructions.

There are many other instructions that take an operand in this way. Some of these are shown in Table 2.

The last instruction is very important for accessing instructions that do not fall into the one-byte-with-operand category listed

Instruction	Hex	Mnemonic	Name
0	0X	j	jump
1	1X	ldlp	load local pointer
2	2X	pfix	prefix
4	4X	ldc	load constant
6	6X	nfix	negative prefix
8	8X	adc	add constant
F	FX	opr	operate

Table 2. Some instructions that take an operand

above. Sixteen instructions are available with no operands: #F0 through #FX; less common instructions are then obtained by using those instructions together with PFIX instructions. All the instructions may be obtained by the use of not more than one PFIX instruction. The full instruction set is very complex and only a few extracts will be given. The floating-point instructions have not been included and neither have the basic arithmetic logic operations, since these are only necessary if you were to write in assembler. Instead, I have concentrated on those instructions that show aspects of the processor which are different from established ones. The opcodes quoted in Table 3 must be converted either in to a prefix followed by an opr or an opr on its own.

The first group of instructions is of interest

Opcode	Mnemonic	Name
22	ldtimer	load timer
2B	tin	timer input
4E	talt	timer alt start
51	taltwt	timer alt wait
07	in	input message
0B	out	output message
0F	outword	output word
0E	outbyte	output byte
0D	startp	start process
03	endp	end process
15	stopp	stop process

Table 3. Opcodes grouped according to the instructions.

because it illustrates the operation of the ALT construct described above; when an ALT construct contains a reference to the timer to wait, it will use those instructions. The second group refers to channels that are used to pass information between processes. The last group is used to control the starting and stopping of processes; stopp is an abnormal end, while endp is

the normal exit from a process.

Transputer applications

For many years the transputer was a solution looking for a problem, because its use was not standard and parallelism was hard to implement and program for. For some reason it was never used, and is still not used in any product, simply as a fast RISC chip. I would have liked to add to the list of applications of transputers by giving one here, perhaps as a programmable arithmetic processor for a PC (Mandelbrot plots in 20 seconds!), but unfortunately the high price tag (£70–£300, depending on speed) and the difficulty of programming them without a proper development board prevented it.

Some of the existing transputer applications are very exciting. One of the newest is a graphics workstation, which, although not cheap (£20,000), is far less expensive than the large mainframe that would have been required without the transputer. It offers a way of manipulating solid shapes, with shading depending on light direction, containing thousands of individual polygons. These can be rotated and zoomed in and out in real time. By contrast, another company produces 3D computer graphics sequences, admittedly on a cinema-size screen, with conventional technology: a Cray supercomputer that takes about an hour to produce one second of film!

Atari are also producing a transputer workstation. This will have a price tag of about £3,000, depending on the number of transputers selected. It can have up to 13, which would give it a performance equal to a present-day mainframe costing about £400,000.

The future
The transputer does seem to be catching on after a long period of stagnation. Inmos moved into profit (about £5 million) in their latest financial year and, if anything, this means that the transputer is here to stay with us!

However, as far as the fairly long-term future is concerned, there are already devices likely to outperform the transputer. Intel have announced recently that they have succeeded in putting a supercomputer style vector processor on to a chip.

Although this may take a long time coming and will not be capable of parallel processing, it will render the transputer obsolete, because it would be able to outperform several hundred transputers. Having said that, it may be as well to be sceptical: perhaps, because of the constraints of putting the vector processor on a chip (preventing high-speed logic being used), it will turn out to have a similar speed to RISC processors around today or RISC processors that have been developed by the time it is ready.

The transputer itself continues to be developed, of course. The T800 is a fairly recent arrival, so that possibly the transputer may have moved another step ahead by the time the competing devices are ready for sale.

References
The two major texts that can be referred to on the transputer are

Transputer Reference Manual
ISBN 0 13 929001 X
Prentice Hall (£19.95)

and

Transputer Instruction Set - a Compiler Writer's Guide
Prentice Hall

There are numerous other technical papers or Inmos documents referred to in the *Transputer Reference Manual*, but these are in the main of interest to specialists only.

Editor's note.
During the final processing stages of this article, Intel announced that later this year it will release a chip, code-named N10, that can perform 150 million instructions per second (MIPs), which makes it about 10 times faster than the fastest transputer available today.

All that is known of the N10 at the moment is that it is manufactured in Intel's 1μ CHMOS process, that its one million transistor design incorporates integer and floating-point add and multiply units on the same die, and that it includes a 4 Kbyte instruction cache and an 8 Kbyte data cache.

DUAL-TONE MULTI-FREQUENCY DECODER

source: Teltone Corporation
design: Robert Krijgsman PE1CHY

The dual-tone multi-frequency (DTMF) dialling scheme for telephone networks was originally developed by Bell Laboratories and introduced in the United States in the mid-1960s as an alternative to pulse dialling. Offering increased speed, improved reliability, and the convenience of end-to-end signalling, DTMF has since been adopted as standard and recommended for use by telecommunications organizations such as CCITT, CEPT, NTTPC, and others around the world. The versatile DTMF decoder described here is a state-of-the art design based on Teltone's single-chip DTMF receiver Type M957.

The DTMF system has been designed to convey dialling information from a subscriber to his local exchange. Since the DTMF keypad is not disabled during the call, the tones may also be sent after the call has been answered. This makes possible a number of interesting applications related to remote signalling and controlling via the telephone network.

Sending DTMF signals is simple with a push-button type telephone, although credit-card like tone senders are available for the older, rotary dialling, telephone sets. After dialling, the miniature loudspeaker in the card is simply held in front of the microphone in the receiver, so that DTMF signals are transmitted to the called subscriber.

Most local telephone exchanges accept calls from both DTMF and pulse-dialling telephones. Especially in areas with relatively few subscribers, however, the local exchange may be compatible with pulse-dialling only. Fortunately, this does not make DTMF-based signalling to the called subscriber impossible because the signals can be sent after establishing the connection by means of pulse-dialling. Special dual-mode telephone sets are available for this purpose: after pulse-dialling, the DTMF keypad is enabled for sending control signals to the called party.

Background to DTMF

When pressed, each of the 16 keys on the DTMF keypad generates a signal composed of one frequency from a high-frequency group, and one frequency from a low-frequency group. The high and low frequency groups are arranged in rows and columns as shown in Fig. 1. Pressing key number 5, for instance, causes the telephone set to transmit two tones simultaneously: one of 770 Hz (row; low-frequency group) and one of 1336 Hz (column; high-frequency group). This 16-digit scheme is often reduced to 12-digit because signals in the 1633 Hz column are reserved for special non-dialling purposes. Many telephone sets in Europe do not have keys A, B, C and D. The DTMF decoder described here supports all 16 codes, however.

The two frequencies that form a DTMF signal are carefully selected for non-harmonic relationship (that is, incidentally, the reason why they sound so ugly).

Signal level considerations

To ensure reasonable communication quality for all subscribers, telephone authorities like British Telecom have

Fig. 1. A DTMF signal is composed of two frequencies selected by pressing a single key on the telephone set.
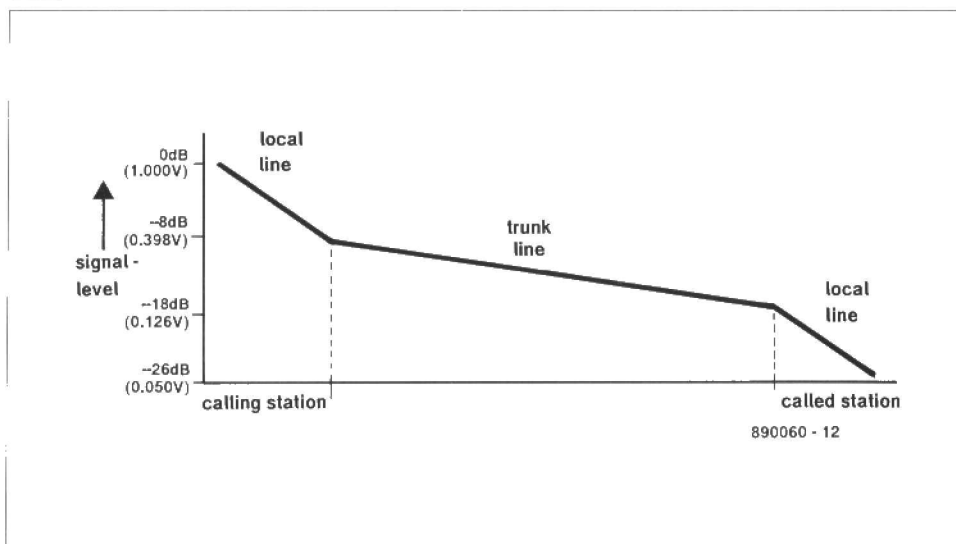
Fig. 2. Graph showing typical attenuation of a signal carried over the telephone network (*courtesy Teltone Corp.*).

drawn up specifications for the maximum signal levels within a particular frequency spectrum transmitted by a telephone set. The same applies to the termination and source impedance, which must be 600 Ω in most countries. A standard telephone set has a transmit level of −11 dBm or 220 mV$_{rms}$ for tones from the low-frequency group, and −9 dBm (275 mV$_{rms}$) for tones from the low-frequency group (0 dBm equals 1 mW into 600 Ω). This simple way of achieving pre-emphasis is required to compensate the high-frequency attenuation introduced by the telephone lines and the amplifiers. A tolerance of 2 dBm is allowed on both transmit levels, which are defined for a nominal termination

impedance of 600 Ω.

Losses inevitably occur as signals are carried from a transmitter to a receiver via the telephone network. Figure 2 shows an example of how telephone lines reduce the effective level of the signal as it is carried from the calling to the called subscriber. From this it will be clear that any DTMF decoder must be able to handle a wide range of input amplitudes. The DTMF decoder described further on in this article meets this requirement by handling signals down to −40 dBm correctly.

Signal attenuation is, unfortunately, not the only problem DTMF receivers have to cope with. Background noise, pulse interference and, of course, voice sig-

nals may be present on the line as shown in the example spectrum of Fig. 3. Because all these signals may contain tone frequencies similar to those representing DTMF digits, high noise immunity and good selectivity are prime design aims for any DTMF receiver.

Teltone's M957 DTMF decoder

The Type M957 is one of a family of DTMF encoder/decoders and related products manufactured by Teltone. The internal schematic of the M957 is shown in Fig. 4. When connected to a telephone line, radio receiver, or other DTMF signal source, the receiver filters out mains-induced noise, dial tones from the exchange, random noise and other interference. It then separates the composite DTMF signal into its high- and low-frequency group components, amplifies and limits these, and digitally measures zero-crossings over averaging periods to achieve digit decoding. With 4 output bits, 16 binary combinations are available, representing codes 1 through 16 or 0 through F hexadecimal.

The only external components required to put the M957 to work are an inexpensive NTSC quartz crystal (3.579 MHz), one resistor and two capacitors. The chip has a sensitive analogue input with a remarkably wide drive margin. The strobe output indicates validity of the data on the 4 digital outputs. The OE (output enable) input allows switching the digital outputs to high-impedance. The strobe output and the OE input make interfacing the decoder to a micropro-
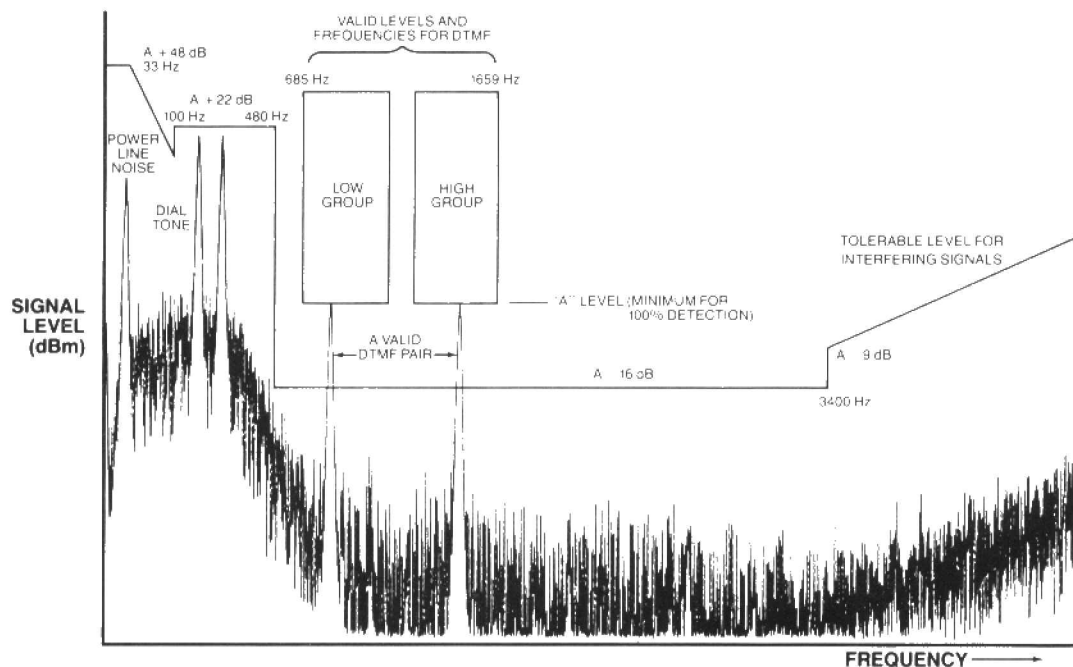
Fig. 3. High noise immunity is a prime design consideration for any dual-tone multi-frequency (DTMF) decoder (*courtesy Teltone Corp.*)

cessor port or bus simple. The logic levels applied to inputs A and B of the M957 define the sensitivity of the analogue input amplifier (see Table 1).

Table 1.

Input sensitivity at Ub=5 V

A	B	min.	typ.	max.	unit
0	0	−32	-	−2	dBm
1	0	−35	-	−5	dBm
0	1	−38	-	−8	dBm
1	1	-	−40	-	dBm

Circuit description

The circuit of the DTMF decoder with digital output interface is shown in Fig. 5. The decoder works automatically: it waits for ring pulses, accepts the incoming call, and then checks the telephone line for valid DTMF signals. The DTMF code sent by the calling subscriber is shown on a 7-segment LED display and made available in binary form on a 4-bit output.

Capacitors C_1, C_5 and C_6 block the direct voltage of about 50 V on the telephone line. Resistor R_9 does not pass current because relay contact re1 is still open. When the subscriber is called, the ring voltage of about 75 V_{pp}/25 Hz causes a current through C_1, R_1 and the LED in opto-coupler IC_7. The phototransistor in the opto-coupler conducts and causes C_2 to be charged slowly via R_2. Two or three ring pulses are required to fully charge C_2. Schmitt-trigger inverters IC_{1a} and IC_{1b} then toggle, so that monostable IC_2 is started via D_2. Relay Ry_1 is energized, LED D_5 lights, and relay contact ry_1 is closed so that R_9 terminates the telephone line: the call is answered.
Meanwhile, C_2 is discharged via $\overline{Q}$ (IC_2) and D_4 to restore the stand-by condition of the ring detection circuit. The calling party can now start sending DTMF codes. When these are received and found valid by the M957, the strobe output, pin 18, goes low briefly, and the decoded digit is fed to the digital outputs and to display driver IC_5 (remember that the digital signal swing is determined by the supply voltage of the decoder chip). The received DTMF digit (0—F) is indicated on a high-brilliance 7-segment display.
Monostable IC_2 keeps the line relay energized because it receives a new trigger pulse from the strobe output via D_3. The relay is de-energized when IC_2 is not triggered within 10 s from the last DTMF code.
Switch S_1 is included to prevent the DTMF decoder automatically accepting incoming calls. The switch thus allows
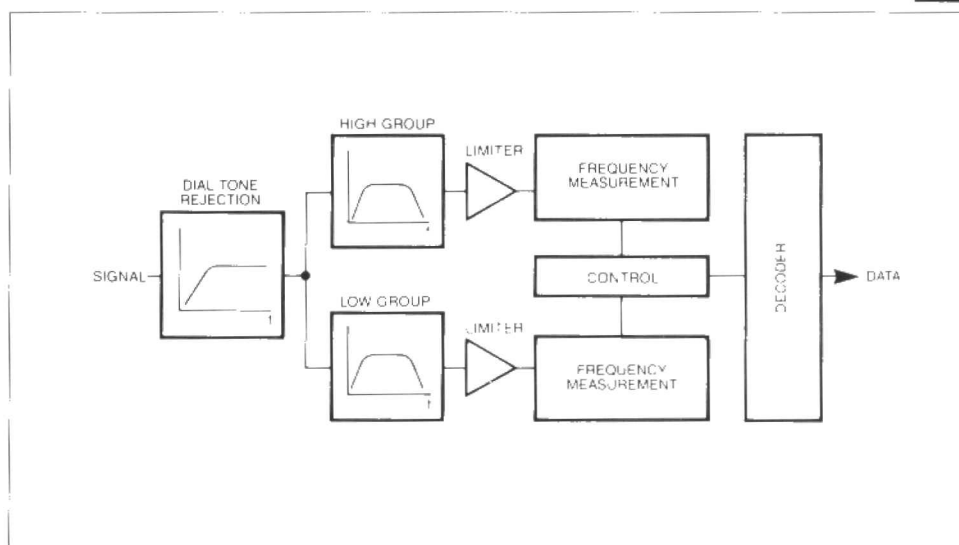
Fig. 4. Internal structure of the M957 single-chip DTMF decoder from Teltone. The analogue signal is converted to a digital code by means of a digital zero-crossing detector (*courtesy Teltone Corp.*).
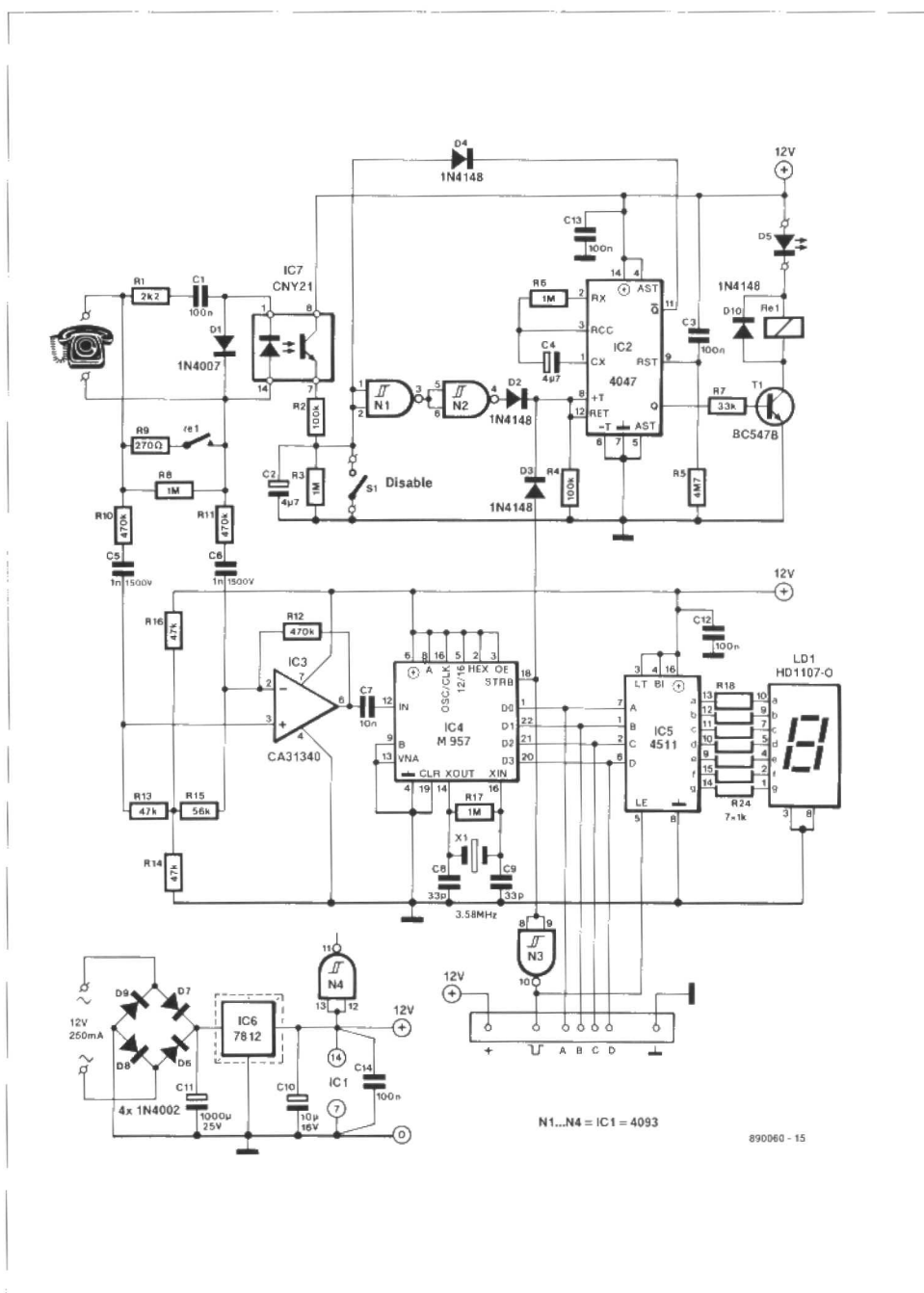
Fig. 5. Circuit diagram of the DTMF decoder with an auto-answering function and digital outputs for remote control applications.
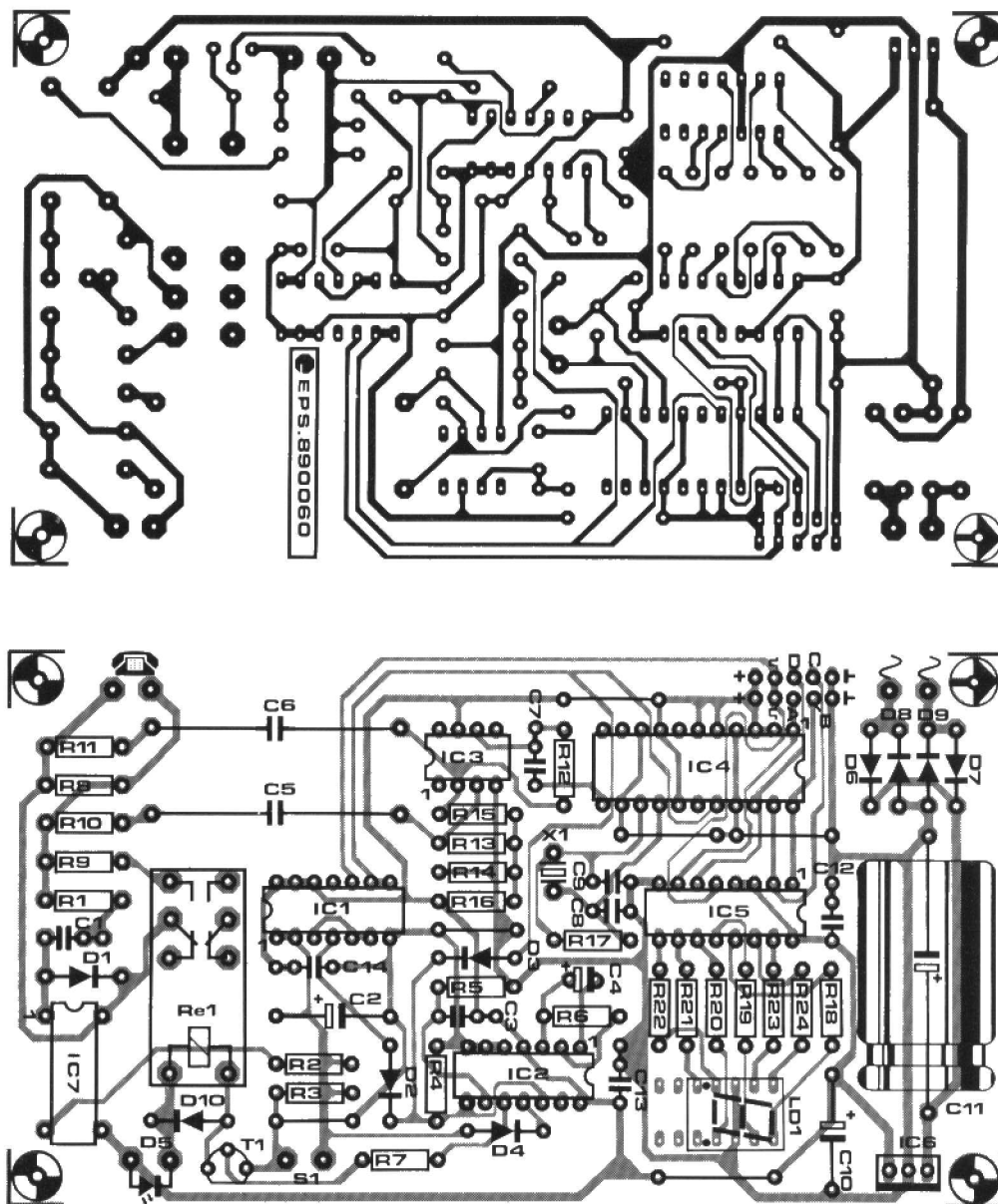
Fig. 6.   Track layout and component mounting plan.

the circuit to be disabled without having to unplug its input cord from the telephone socket.

Construction

The printed-circuit board shown in Fig. 6 greatly facilitates the construction of the DTMF decoder. All the parts are simply fitted in accordance with the overlay on the board and the Parts List. The decoder is conveniently powered from a 12 VAC/250 mA mains adaptor. Voltage regulator IC_6 must be fitted with a small TO-220 style heat-sink. Alternatively, it may be bolted to the enclosure panel with an insulating washer in between. Capacitors C_5 and C_6 must have the specified voltage rating of 1500 VDC to prevent any chance of an

Parts list

Resistors (±5%):
R1=2K2
R2=100K
R3;R6;R8;R17=1M0
R4=100K
R5=4M7
R7=33K
R9=270R
R10;R11;R12=470K
R13;R14;R16=47K
R15=56K
R18...R24 incl.=1K0

Capacitors:
C1;C3;C12;C13;C14=100n
C2=4µ7; 63 V; axial

C4=4µ7; 63 V; radial
C5;C6=1n0 1500 V
C7=10n
C8;C9=33p
C10=10µ; 16 V
C11=1000µ; 25 V

Semiconductors:
D1=1N4007
D2;D3;D4;D10=1N4148
D5= red LED
D6...D9 incl.=1N4002
T1=BC547B
IC1=4093
IC2=4047
IC3=CA3140
IC4=M957 (Teltone)
IC5=4511
IC6=7812
IC7=CNY21 (listed by

Universal Semiconductor Devices Ltd.)

Miscellaneous:
S1= miniature SPST switch.
Re1= 12 V PCB-mount relay; eg. Siemens V23037-A0002-A101.
X1= quartz crystal 3.579 MHz (Cricklewood Electronics Ltd).
LD1= HD1107-O (Siemens; ElectroValue).
TO-220 style heat-sink for IC6.
PCB Type 890060 (see Readers Services page).

dialing access and control.

Teltone Corporation ● P.O. Box 657 ● 10801 120th Ave. ● N.E. Kirkland WA ● USA 98033-0657. Telephone: (206) 827-9626.

Teltone Corporation ● 7A Hill Avenue ● AMERSHAM HP6 5BD. Telephone: (0494) 728099.

UK distributor is Chesilvale Ltd. ● 10 Woodland Road ● CLIFTON BS8 1HQ. Telephone: (0272) 736166. Fax: (0272) 736516.

Note: the DTMF decoder described here is not type-approved by British Telecom and may not be connected to the public switched telephone network (PSTN). In countries other than the UK, the relevant PTT authorities should be contacted for information on type-approval.

Fig. 7. The completed DTMF decoder is a compact and reliable unit. The 4-bit strobed output simplifies interfacing to a wide range of digital equipment, including PCs and other home computers.

external voltage being applied to the telephone line. Along the same lines, the stated optocoupler meets the requirements for minimum insulating voltage. Digital output signals A—D and the strobe signal are available on a 10-way pin header and may be used for driving a digital circuit via a short flat-ribbon cable. An external level shifter such as the Type 4050 is required to obtain TTL-compatible pulses. It is, however, also possible to feed the decoder from a 5 V supply, obviating the need for a level shifter.

Applications

The DTMF decoder provides reliable code transmission via many media, whether telephone lines, local control networks, industrial wiring systems, or radio equipment. Computer owners may ring up their own system at home and turn the mains to the machine on with a particular DTMF code that controls a relay. The AUTOEXEC.BAT file may then start a particular programme, whose course is determined by bit-combinations supplied by the DTMF decoder. One code must be reserved, however, to shut the system down.

Remote control of coffee machines, cassette and video tape recorders and other equipment is also possible by reserving two codes to turn each device on and off. The LED display always shows the last received code.

The remote control system can be given some security by fitting non-standard, but matching, quartz crystals in the DTMF sender and decoder. In the proto-

type, PAL crystals (4.433 MHz) fitted in a DTMF sender and the decoder described here resulted in acceptable rejection of the standard (3.579 MHz derived) tones.

Other applications of the DTMF decoder include selective calling for mobile radios and tape control for slide presentations.

Source: Teltone Application Guide SC1: *Applications for DTMF and pulse*

APPLICATION NOTES

The contents of this article are based on information obtained from manufacturers in the electronics industry, or their representatives, and do not imply practical experience by *Elektor Electronics* or its consultants.

GET OFF THE BUS WITH TAXICHIP™ DEVICES

The Am7968 TAXIchip™ transmitter and Am7969 TAXIchip receiver chip set from Advanced Micro Devices is a general-purpose interface for very high-speed (4—12.5 Mbytes/s) point-to-point communications over coaxial or fibre-optic media.

In the past, high-speed data transfers required the use of parallel buses. Sending lots of bits simultaneously was faster than sending them one-bit-at-a-time. But this solution was messy: multiple data paths added cost and bulk to a system. Performance issues cropped up, too. Parallel buses were only practical over relatively short distances; transmission speed was limited by cross-talk and radio interference (RFI) considerations. Also, reliability suffered as the number of wires and connector pins increased.

Advanced Micro Devices (AMD) has developed TAXIchip™ (*T*ransparent *A*synchronous *X*mitter-receiver *I*nterface) devices that provide the performance of multiple-path design without all the drawbacks. Serial communication links are easy with the high-speed Am7968/Am7969 chip set. These ICs enable communica-tion engineers and designers to replace parallel buses with a serial link, while maintaining the performance of the old parallel system. The bulk, weight and cost of parallel ribbon cables and associated connectors are eliminated without reducing data throughput and without changing the system architecture.
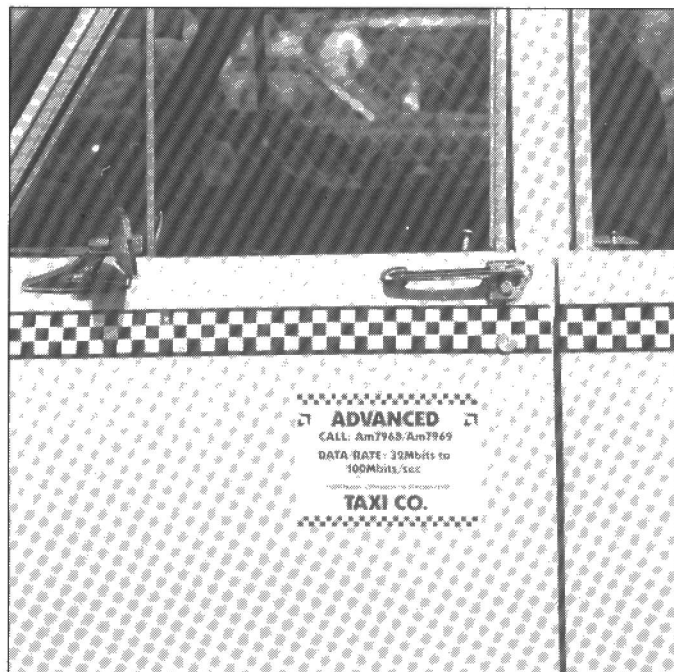
Speed up data communications...

The 100 Mbps effective data throughput rate of the TAXIchip set is 10 times faster than the data rate of conventional RS-422 line drivers and receivers — the fastest point-to-point communications standard in existence today.

The versatility of the Am7968/Am7969 chip set allows top performance to be achieved with lower hardware cost and minimal

GET OFF THE BUS
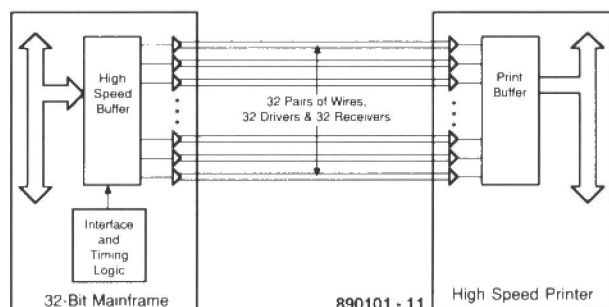
WITH TAXIchip DEVICES
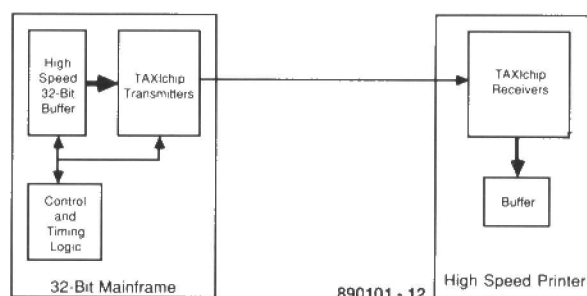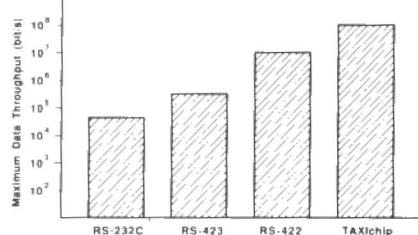
The TAXIchip set offers a data throughput rate higher than any of the currently used interface standards.

be programmed to select one of three bit modes of operation:

- 8 data and 4 command bits;
- 9 data and 3 command bits;
- 10 data and 2 command bits.

For wider parallel inputs, TAXIs can be cascaded to squeeze multiple-byte words into a single serial link. Flexible byte-width and the option to cascade TAXI chips give freedom to move a maximum amount of data with a minimum number of components.

The TAXIchip transmitter

The AM7968 transmitter has 12 parallel inputs, segregated into data and command lines. Normal message traffic is sent over data lines. Supervisory control
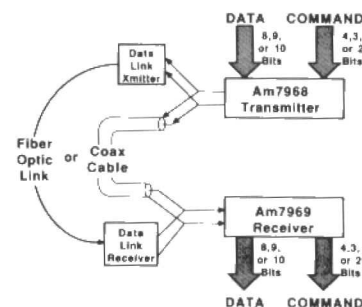
Typical basic applications of the TAXIchip set from AMD.

RFI effects. Coaxial cable can be can be driven direct by TAXIchip products, and is the lowest-cost media for short-to-moderate distances. Optical Data Links (ODL) are also simple to interface to the chip set, affording all the advantages of fibre-optic communication. These benefits include long-distance coverage, immunity to Electromagnetic Interference (EMI) and RFI, wide bandwidth and electrical insulation.

...and increase communications distance

As Am7968/Am7969-based designs communicate over a serial link, distances are not limited by the same constraints that plague parallel buses. Cross-talk and bit-to-bit skew are not problems with TAXI-based connections. And since the Am7968 and Am7969 were designed specifically to interface with high-speed optical data links, fibre optic cable may be used to cover distances up to several miles.

Accommodates all bus widths

The Am7968 TAXIchip transmitter and Am7969 TAXIchip receiver form a flexible interface solution for high-speed point-to-point data communications. The TAXIchip set operates much like a single parallel register. Data is loaded into one side and read from the other, which is separated by a long serial link. TAXIs have 12 parallel interface pins designated as either Data or Command bits. Data represents the normal message traffic between host systems. Commands may originate from the communication control section of a host; they occur at a relatively infrequent rate but have priority over data. Each TAXI can

TAXIchip devices

- 10 times faster than conventional RS-422
- Data throughput up to 100 Mbps (12.5 Mbytes/sec)
- Direct drive of coaxial cables, or interface with fibre-optic links
- Programmable byte width
- Two priority levels for information transfer
- Cascadable, for buses 16, 32 or more bits wide
- Complete on-chip PLL and crystal oscillator
- Simple strobe/acknowledge handshake
- Straight-forward TTL parallel bus interface
- Single +5 V supply

information is handled by command lines. The DATA MODE SELECT pin may be tied low, high or left open to determine the mix of data and command pins. The transmitter sends data if all of the command lines are logic 0 when the device is strobed; command information is sent if any of the command lines are logic 1 when the transmitter is strobed.

Parallel information is captured in the input latch on the rising edge of a strobe pulse. An acknowledge signal is then sent back from the transmitter when it is ready to accept the next byte. Information is transferred to the encoder latch first and then through the data encoder and into the shift register, on two successive byte clocks. Encoded patterns then come out of the shifter and through
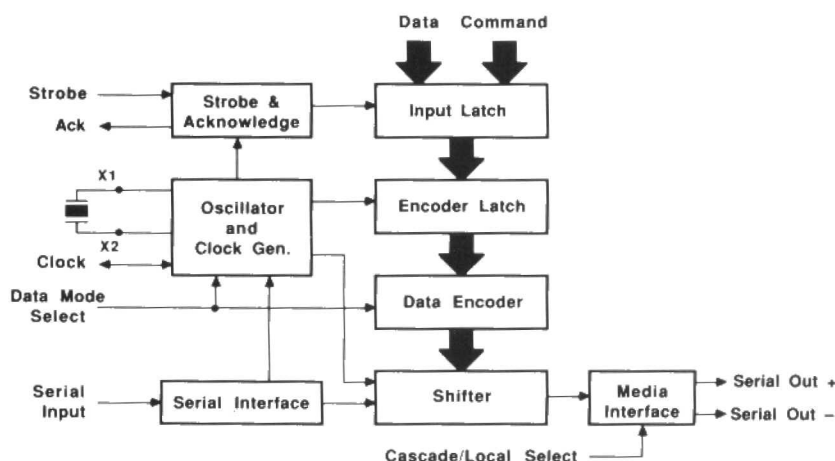
Internal block schematic of the Am7968 AXIchip transmitter

the media interface one bit at a time. Voids between transmitted data or command bytes are filled with special sync patterns which keep the Am7969 PLL synchronized, and define byte boundaries. As data does not have to be sent every byte cycle, parallel data throughput in bytes/s can be slower than the minimum specified transmission rate (40 MHz; 32 Mbps) would seem to indicate.

The data rate is set indirectly by an external quartz crystal which oscillates at the byte rate. An internal PLL multiplies the byte rate by 10, 11 or 12 (depending on whether the TAXIs are programmed to be in 8, 9 or 10-bit mode) to arrive at the serial transmission rate. TAXIs use 4B/5B and/or 5B/6B encoding schemes, essentially expanding each byte by two
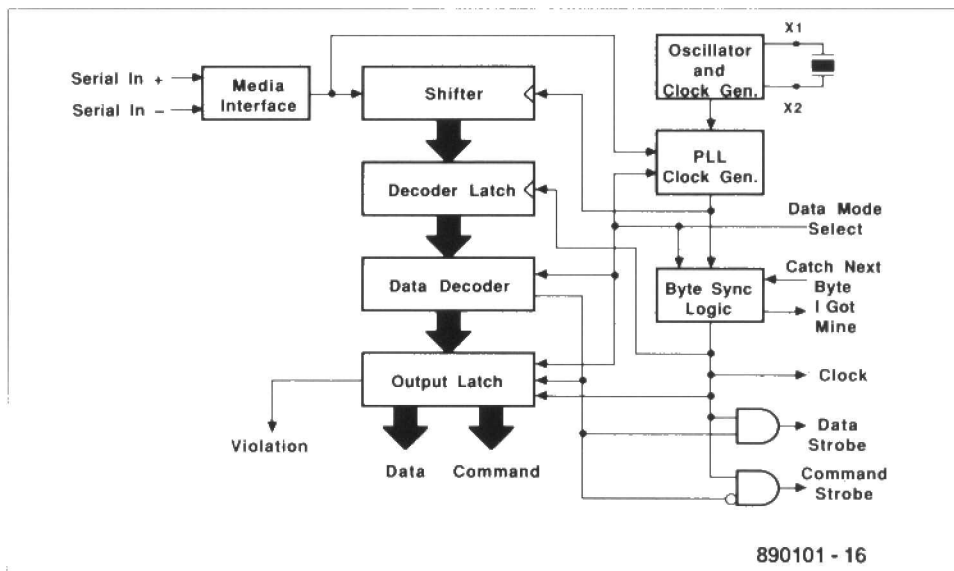
Internal block schematic of the Am7969 TAXIchip receiver.

bits.

The SERIAL INPUT pin allows transmitters to be wired in parallel to funnel very wide data paths into one high-speed channel.

The TAXIchip receiver

During reception, the Am7969 accepts message patterns into its shifter. In the next byte clock-cycle, the pattern is parallel-loaded into the decoder latch. One clock cycle later, the pattern is passed through the data decoder, recognized as either data or command, and captured by the appropriate output latch. A new data or command byte is then issued from the latch and accompanied by a data or command strobe. The byte-rate clock is also extracted from the incoming patterns and provided for use by the receiving system.

If any of the received patterns have been corrupted into invalid 4B/5B or 5B/6B

codes, then the VIOLATION pin will identify the coding violation.

Handshaking signals CATCH NEXT BYTE and I GOT MINE are handshaking signals which allow receivers to be cascaded.

High-speed data communication design made easy

The TAXIchip set directly drives coaxial cable to achieve very fast and cost-effective point-to-point communication. Or, fibre-optic cable and ODLs may be used if higher speed, longer distance, or electrical insulation is required.
Fibre-optic data communication is coming of age. High bandwidth and low signal attenuation are attractive characteristics of fibre-optic cable. Optical data transmission also offers noise immunity, minimizes RFI emission, eliminates ground loops and provides data security. Component costs are fall-

ing to a point where these advantages are quickly becoming cost-effective for a wide range of applications.

Cascadable for wider data paths

Multiple TAXIchip sets can be wired in parallel to share a single serial channel. This allows 16- or 32-bit wide buses to be replaced by a single wire, substantially cutting interconnection cost in systems with many communications paths or long cable runs.

The Am7968 can be set to local or cascade mode by tying a single pin, CASCADE/LOCAL SELECT, high or low. Local mode allows a single transmitter to send messages over a serial channel. Additional transmitters can share this channel, running in parallel with the primary Am7968, if they are put in cascade mode. All cascaded transmitters simultaneously latch a parallel word from the data or command inputs upon seeing a common strobe pulse. Each byte of this word is individually encoded and then shifted one bit at a time, through the chain of transmitters and on to the media. Since the entire word passes through the primary transmitter, the maximum transmission rate of this configuration is still 125 MHz, with a maximum data throughput of 100 Mbps.
The Am7969 has two pins, CATCH NEXT BYTE and I GOT MINE, which can be configured to allow TAXIchip receivers to operate singly or in parallel. These handshake signals tell each Am7969 when to capture its respective byte from the media, so that multiple-byte words can be reconstructed correctly on the receiving end of a communications channel. These wiring options allow TAXI-based
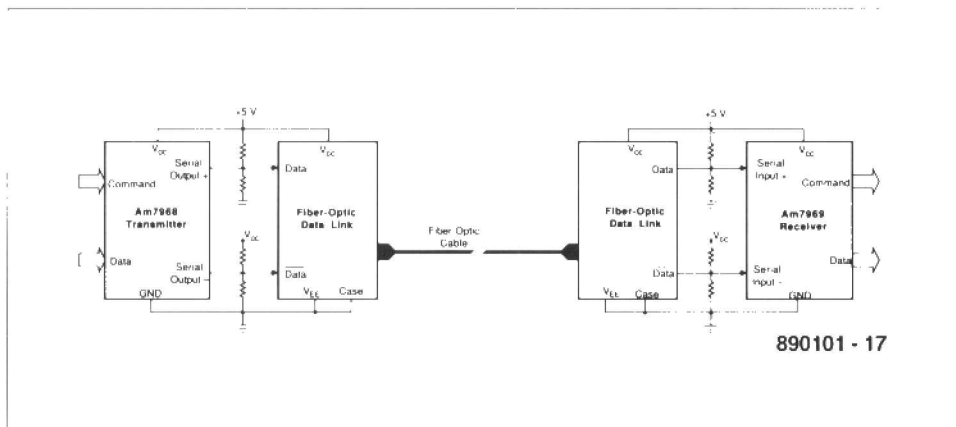
The only additional hardware needed to connect the TAXIchip set to an optical fibre network are a fibre-optic data link and a simple resistance network, which supplies bias voltage and terminates the transmission line.

Am7968

DIPs   PLCC*

Am7969

DIPs   PLCC*
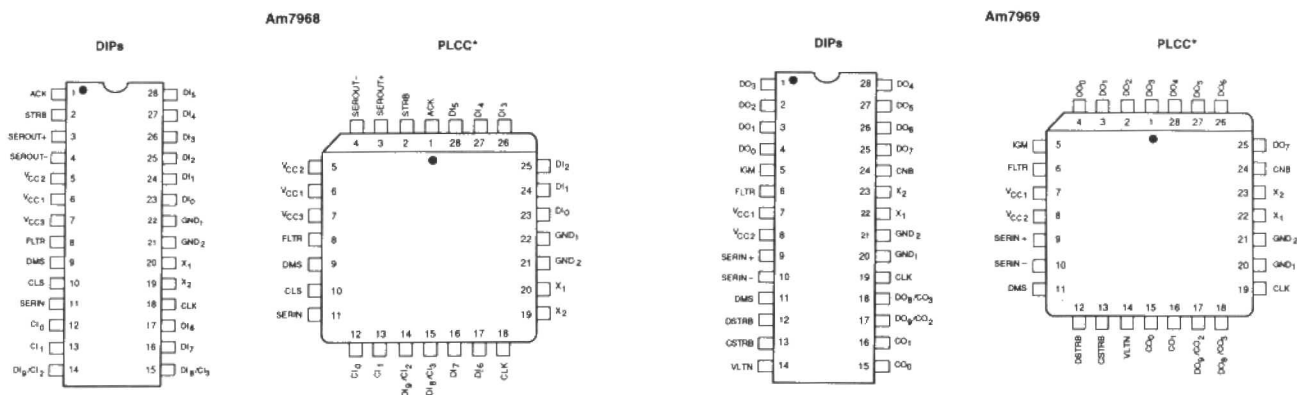
communication channels to be structured in a number of ways. For example, one transmitter can broadcast a message to several receivers, multiple TAXIchip pairs can be cascaded to share a serial channel, or, for the fastest data throughput, multiple Am7968 and Am7969 TAXIchip pairs can operate in parallel, each pair using a separate serial link.

Simple parallel interface

TAXI's bus interface lines operate much like those of an ordinary register. A simple strobe/acknowledge handshake is used to latch parallel information into the Am7968 TAXI transmitter. Data is then automatically — transparently —

encoded, serialized, de-serialized, decoded, and issued from the Am7969 TAXIchip receiver with an accompanying strobe pulse. In a system, these two chips act much like a single latch or register.

Integrated 125 MHz phase locked loop

Am7969 TAXIchip receivers are claimed to have the fastest commercially available integrated circuit. Using so-called *Current Mode Logic* technology, the PLL operates over a frequency range of 40 to 125 MHz, and will track data with jitter of up to 40% of a bit time. The PLL requires only a single capacitor

with ±10% tolerance to be made operational.

Group encoding and identification of code violation

4B/5B and 5B/6B encoding schemes are used to combine clock and data into a single bit-stream and ensure that data integrity is maintained. These encoding methods essentially expand each byte by two bits. The resulting patterns are relatively free of DC bias and can pass through a standard AC-coupling circuit with minimal distortion.

Since these coding methods are so efficient, bandwidth requirements for the serial link are held to a minimum. The 4B/5B code uses a 125 MHz signal to transfer data at a rate of 100 Mbps. The popular Manchester encoding/decoding scheme would require a 200 MHz signal to achieve the same 100 Mbps data throughput.

The Am7969 detects most noise-induced transmission errors. Invalid 4B/5B coding patterns are recognized and identified on a special VIOLATION output pin.

TAXIchip™ is a trademark of Advanced Micro Devices, Inc.

Further information on the Am7968/Am7969 may be obtained from **Advanced Micro Devices** • P.O. Box 3453 • Sunnyvale CA 95088 • USA.

AMD UK can be contacted at (0925) 828008 (Manchester area) or (04862) 22121 (London area).

Am7968/7969 TAXI Chip Set
Transparent Asynchronous Transmitter/Receiver Interface

Parallel Data

Data Link Transmitter

Am7968 Transmitter

Fiber Optic Link or Coaxial Cable

Data Link Receiver

Am7969 Receiver

Parallel Data

CODE CONVERTOR FOR CENTRONICS-COMPATIBLE PRINTERS
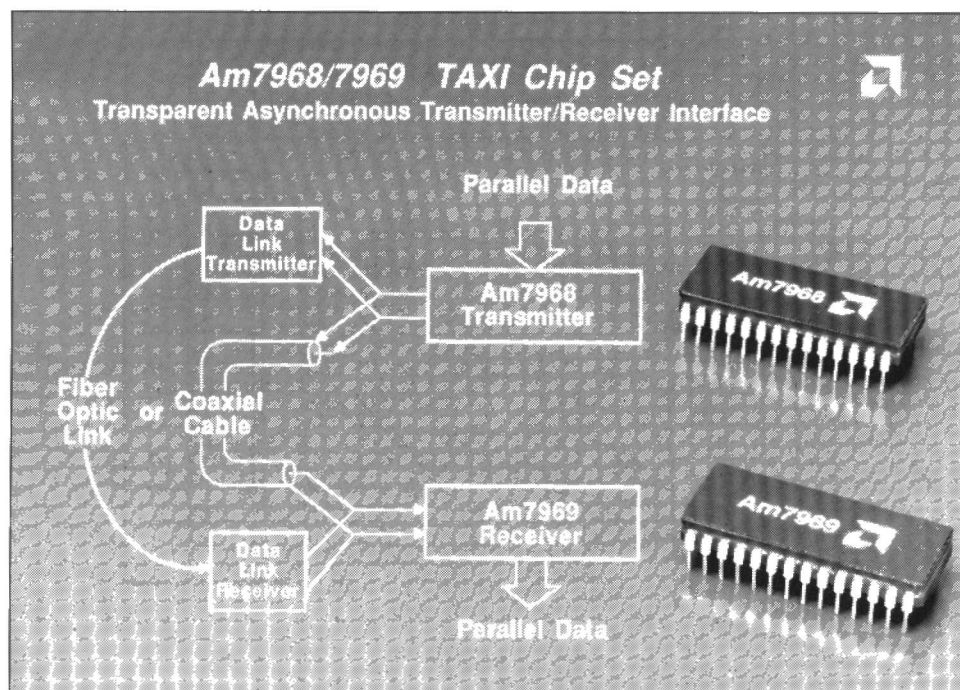
by N. Willmann

This simple EPROM-based converter solves awkward interfacing problems raised by incompatibility of the character codes used by certain computer programs and those actually available in parallel printers.

Very few problems can be expected to arise when a computer sends nothing but straight ASCII to a Centronics-compatible printer. Not so with many special signs, diacriticals and Greek characters, however, which many printers have available but never seem to get right until lots of DIP switches have been reset, and the printer is found to run in a mode entirely different from that desired.

Fortunately, today's wordprocessors provide configurable printer drivers that obviate the use of copied ASCII code tables and overviews of DIP switch configurations. Such a printer driver is, however, not available with many other programs, such as assemblers, interpreters, compilers (BASIC, PASCAL), CAD/CAM programs (netlisting!) and a number of databases. Users of these programs are often faced with real difficulties when characters beyond the alphabet, numbers 0 to 9, and the basic set of arithmetic operators are to be used on an otherwise perfectly functioning printer. Not surprisingly, the first hardcopy produced on the fly looks far worse than that of the wordprocessor.

The above deficiency is, in general, simpler to overcome with hardware than with software, which, in view of the required speed, must needs rely on machine code and code conversion tables. In many cases, code-converting hardware is available from printer manufacturers in the form of a ROM claimed to make the printer compatible with a particular type of computer. The main disadvantages of such a chip are first and foremost that it is often relatively expensive, and secondly that it is difficult to remove and replace when, occasionally, another computer is used. A fair compromise between printer driver software and the printer-resident
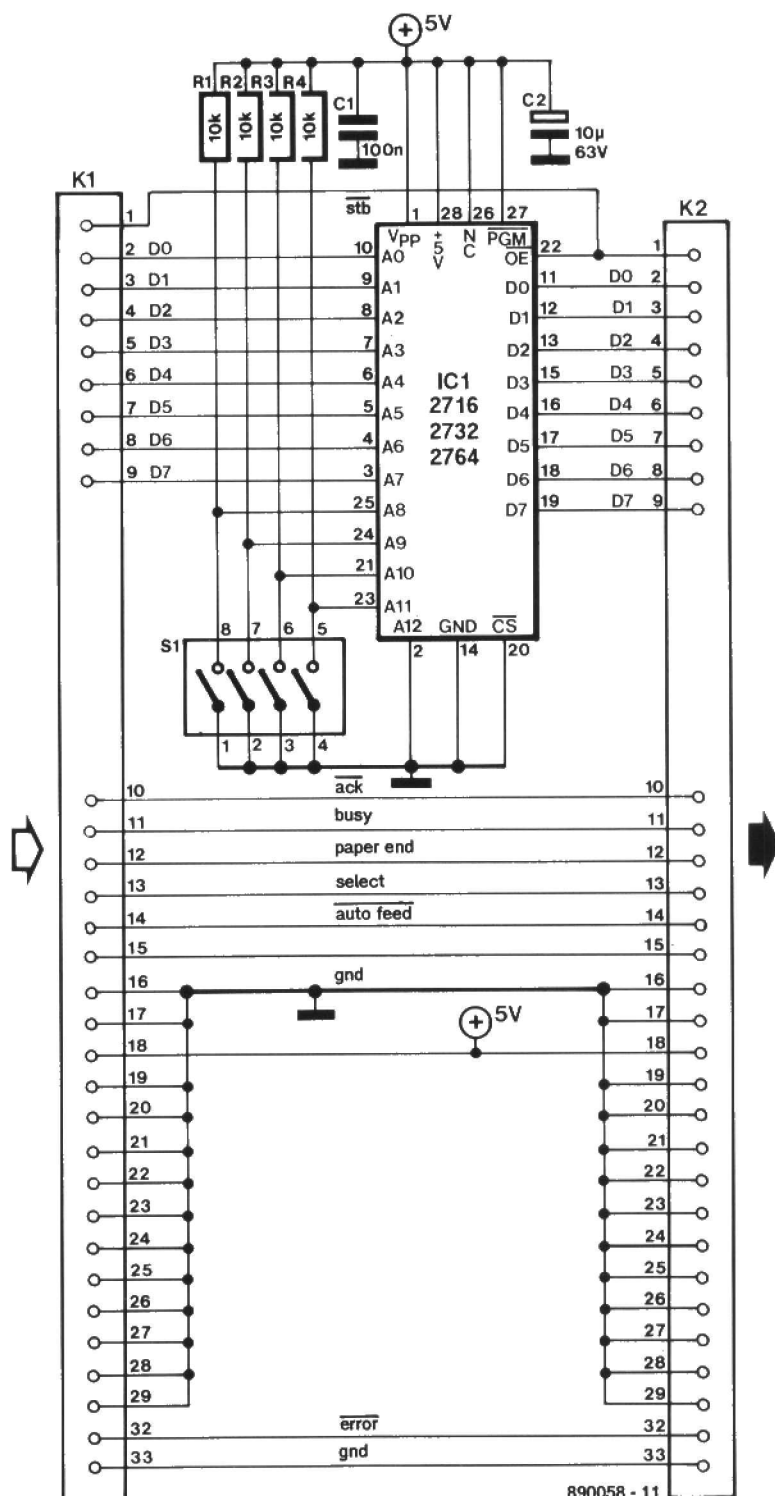
Fig. 1.   Circuit diagram of the code convertor for Centronics-compatible printers.

Table 1.

address	data															
x00...x0F	00	01	02	03	04	05	06	07	08	09	0A	0B	0C	0D	0E	0F
x10...x1F	10	11	12	13	14	15	16	17	18	19	1A	1B	1C	1D	1E	1F
x20...x2F	20	21	22	23	24	25	26	27	28	29	2A	2B	2C	2D	2E	2F
x30...x3F	30	31	32	33	34	35	36	37	38	39	3A	3B	3C	3D	3E	3F
x40...x4F	40	41	42	43	44	45	46	47	48	49	4A	4B	4C	4D	4E	4F
x50...x5F	50	51	52	53	54	55	56	57	58	59	5A	5B	5C	5D	5E	5F
x60...x6F	60	61	62	63	64	65	66	67	68	69	6A	6B	6C	6D	6E	6F
x70...x7F	70	71	72	73	74	75	76	77	78	79	7A	7B	7C	7D	7E	7F
x80...x8F	80	81	82	83	84	85	86	87	88	89	8A	8B	8C	8D	8E	8F
x90...x9F	90	91	92	93	94	95	96	97	98	99	9A	9B	9C	9D	9E	9F
xA0...xAF	A0	A1	A2	A3	A4	A5	A6	A7	A8	A9	AA	AB	AC	AD	AE	AF
xB0...xBF	B0	B1	B2	B3	B4	B5	B6	B7	B8	B9	BA	BB	BC	BD	BE	BF
xC0...xCF	C0	C1	C2	C3	C4	C5	C6	C7	C8	C9	CA	CB	CC	CD	CE	CF
xD0...xDF	D0	D1	D2	D3	D4	D5	D6	D7	D8	D9	DA	DB	DC	DD	DE	DF
xE0...xEF	E0	E1	E2	E3	E4	E5	E6	E7	E8	E9	EA	EB	EC	ED	EE	EF
xF0...xFF	F0	F1	F2	F3	F4	F5	F6	F7	F8	F9	FA	FB	FC	FD	FE	FF

ROM is offered by the circuit of Fig. 1. This code converter is EPROM-based and allows any printer code to be replaced by another. Due care should be taken, however, to avoid the conversion of codes related to printer control commands (escape sequences).

Circuit description

The circuit works on a simple basis. The databyte received from the computer's Centronics port is used to address an EPROM programmed to hold the desired printer codes. The EPROM thus contains a list with 256 (0-255) entries that supply a particular printer code for each computer code applied. The EPROM may contain several such lists, which are selected by switch S_1. The maximum number of lists is determined by the EPROM used: a 2 Kbyte Type 2716 may hold up to 8 lists, while the larger Types 2732 (8 KByte) and 2764 (16 KByte) have capacity for 16 lists (the 2764 then has further free memory, which is not used here). The choice of the EPROM for the present project is mainly governed by availability and price (the 2764 is probably the cheapest these days).

The larger part of the circuit is formed by direct connections that carry the printer control and handshaking signals between the input connector, K_1, and the output connector, K_2. Only the 8 datalines are 'broken' and applied to the 8 least-significant address lines, A_0 through A_7, of EPROM IC_1. The $\overline{STB}$ (strobe) line is tapped to drive the $\overline{OE}$ (output enable) pin on the EPROM. This is done to ensure that the data on the 8 output lines is stable when it is loaded by the printer.

The interface is powered from the +5 V output available at pin 18 of the printer. It should be noted that not all printers have such an output, which may, therefore, have to be provided by the user. Alternatively, the code convertor may be powered from a small, regulated, 5 V
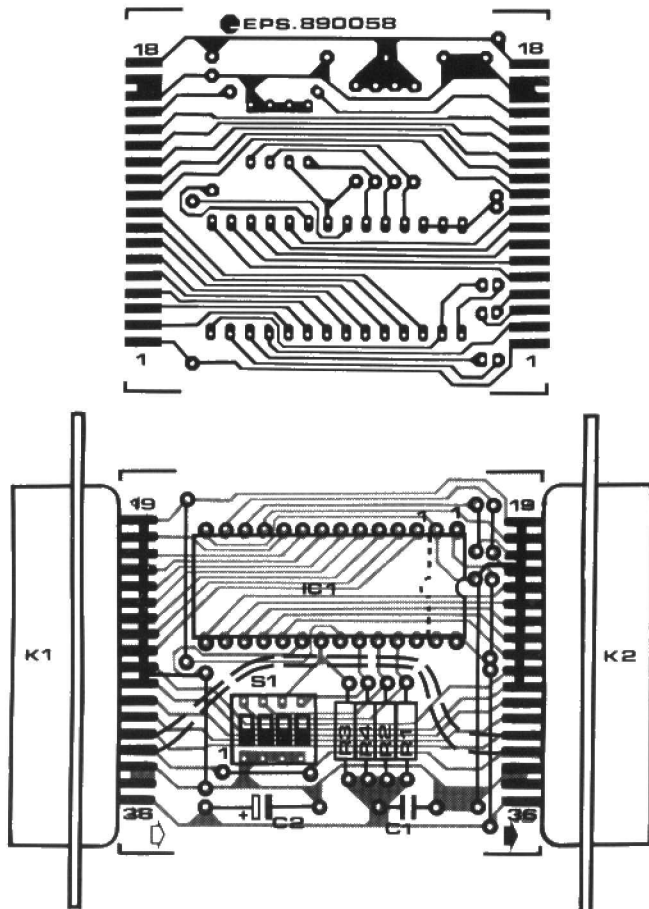
Fig. 2. The contact fingers at the PCB edges can be inserted between the two rows of pins on the Centronics connectors.

Parts list

Resistors (±5%):
R1...R4 incl.=10K

Capacitors:
C1=100n
C2=10μ; 63 V

Semiconductor:
IC1=2764 (see text)

Miscellaneous:
S1= 4-way DIP switch block.
K1= female 36-way Centronics connector with straight pins.
K2= male 36-way Centronics connector with straight pins.
PCB Type 890058 (not available through the Readers Services).

source. The 5 V supply rail of the circuit is made available on K₁ so that it may power further equipment, such as a printer buffer.

Conversion list

The data given in Table 1 may appear useless at first sight because it leaves applied data unchanged. The table is purposely compiled as it is, however, for two reasons.

First, it is the list used in the lowest address range of the EPROM (list number 1, EPROM address range 0000ʜ-00FFʜ or 0ᴅ-225ᴅ). Because the table does not alter the computer codes, it guarantees correct operation of the printer when this is used in graphics mode.

Secondly, the given data forms the starting point for the building of other lists in the EPROM. Copies of the default list are then ready for editing, i.e., entering one's own codes, on paper. To do this, simply look up databyte at the address that corresponds to each code you want to change, eradicate the given databyte, and replace it with the required code. Now load or type the new list into your EPROM programmer, starting at EPROM address 0100ʜ or 256ᴅ (list number 2). The other tables, when required, are built and loaded in a similar manner. Do not forget to document the settings of S₁ for the available conversion tables.
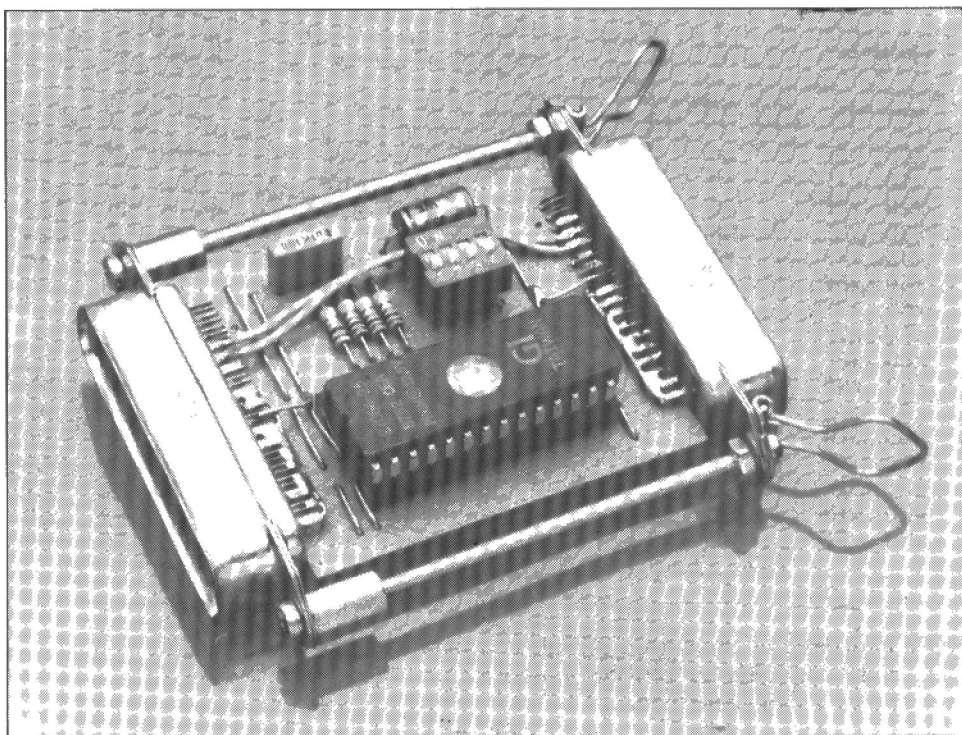
Construction

The printed-circuit board shown in Fig. 2 is designed such that pins 1 through 18 of Centronics-style connectors K₁ and K₂ can be soldered direct on to the contact fingers at the edges. The other row of pins, 19 through 36, is wired in accordance with the overlay on the board. DIP switch block S₁ may be mounted at the copper or component side of the board, whichever location is most convenient for easy accessing once the completed board has been installed in an enclosure.

Award for young radio amateur of the year

The Department of Trade and Industry (DTI) recently announced its sponsorship of the Young Amateur of the Year Award for 1989's outstanding achievement by a young amateur radio enthusiast.

Anyone who is under 18 and:
- is keen on do-it-yourself radio construction; or
- is interested in using radio and gaining operating skills; or
- is using radio for a community service, such as helping the disabled or in emergency communication networks; or
- is good at encouraging interest in amateur radio; or
- is involved in amateur radio in any way, such as in a school scientific project

is eligible for the 1989 Award and its £250 prize.

The prize for the most outstanding achievement between 1 April 1988 and 31 July 1989 will be awarded by the DTI and presented at the Radio Society of Great Britain's HF Convention in October 1989.

On top of the £250 prize, DTI will provide every genuine entrant with a copy of its coloured chart of radio frequency allocations in the UK. The winner will also get to see DTI's radio experts at work at its Radio Monitoring Station at Baldock in Hertfordshire.

Last year's Award went to 15-year-old Andrew Keeble, from Norwich, for his enthusiasm in encouraging others' interest in radio, his radio construction skills, and voluntary activities. He was presented with his prize by HRH Prince Philip who is patron of the RSGB. Andrew also saw the work of the Baldock Station, and in addition got some surprise awards which were: a one-week training course at the College of Marine Electronics, sponsored by the Mobile Radio Users' Association; an engraved RC14 receiver given by the RSGB; and a week in Vienna as guest of the OVSV, the Austrian national radio society.

How to enter the competion

Closing date for applications is 31 July 1989. Entrants do not need to be a radio licence holder to enter, and the competition is open to anyone in the UK, the Channel Islands, or the Isle of Man, who is under 18 on 31 July.

Young people may either enter directly, or an adult may nominate a candidate for the Award on the strength of the adult's knowledge of the entrant. In the case of the candidate making his own entry, it should be supported by a short letter from a licenced radio amateur or other adult who knows him or her well and vouching for these achievements.

Entries should be reasonably full and give a clear picture of the candidate's achievements. Enough detail should be given to allow an assessment to be made of their quality.

Candidates should submit some background information on themselves with their entry, e.g., their other hobbies, especially if amateur radio is involved, their school achievements and, if known, their career intentions.

Applications or nominations for the Award must be sent to:

The Secretary ● Radio Society of Great Britain ● Lambda House ● Cranborne Road ● Potters Bar ● Hertfordshire EN6 3JE. Telephone: (0707) 59015.

SUPER-VHS-TO-RGB-CONVERTER

design: ELV GmbH
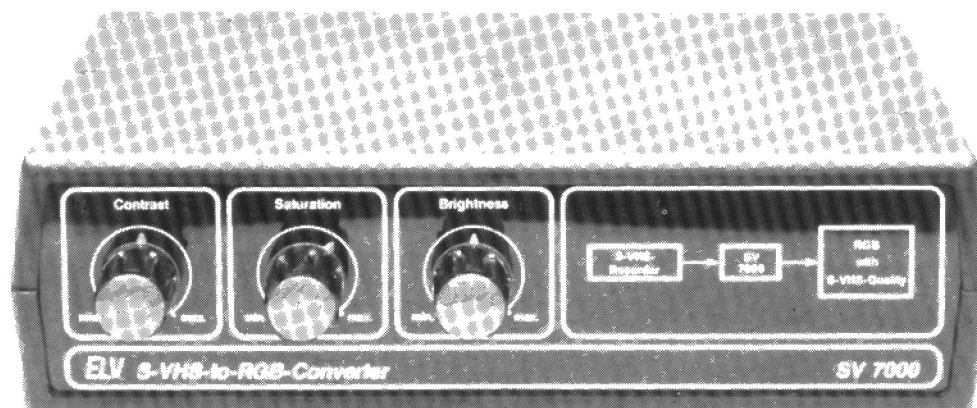
Almost any radio and TV salesman will confidently tell you that the excellent picture quality from a S-(super) VHS video cassette recorder can only come into its own on a new, compatible, TV set with separate luminance and chrominance inputs. Unfortunately, these TV sets are still relatively rare and also quite expensive. The SV7000 converter presented here is a cost-effective alternative to a new TV set. This is because it provides a function not available on most S-VHS VCRs: conversion of the luminance and chrominance signals into RGB signals that can be applied to a SCART-compatible TV set.

There can be no argument that super-VHS (S-VHS) video tape recorders provide higher resolution and better picture quality than any of the currently available high-quality (HQ-) models. Although the price of S-VHS VCRs is coming down, and many video enthusiasts are beginning to see the real benefits of the new system, the cost of a matching TV set is nearly always a prohibitive factor. For some reason, S-VHS compatible TV sets are in rather short supply, so that many owners of a new recorder are forced, for the moment, to make do with the reduced picture quality offered by their existing TV set, which may be upgraded in some cases with an S-VHS interface. Alternatively, an audio/video (AV) connection, or, worst of all, a UHF modulator, may be used to get the new S-VHS VCR to work with a current TV set. Clearly, these solutions give poor results considering the outlay for the new VCR.

The SV7000 S-VHS-to-RGB converter described here accepts the luminance and chrominance signals from the S-VHS VCR, and converts these into RGB signals for driving the corresponding in-

puts on the SCART input of the TV set. This solution is very cost-effective compared with buying an S-VHS compatible TV set, and guarantees the best possible picture quality from the combination of an S-VHS recorder and a SCART-compatible TV set.

Simple to connect

Most currently available S-VHS VCRs have a so-called mini-DIN 4-way socket that carries the two components of the picture signal, luminance (Y, brightness information) and chrominance (C, or colour information). These two signals are fed to the corresponding inputs on an S-VHS TV set via a flexible, 2-way, individually screened, cable terminated in 4-way mini-DIN plugs.

The SV7000 is inserted between the Y-C output of the S-VHS recorder and the SCART input of the TV set. The audio signal(s) from the recorder is also applied to the SV7000, in this case via a pair of phono plugs and mating sockets. The SCART output of the SV7000 carries all the necessary signals for driving the corresponding input on the TV set: colour signals R (red), G (green) and B (blue), composite synchronization and

the audio signal (mono VCR) or signals (left/right; stereo VCR). Hence, the SV7000 is connected to the TV set via a standard SCART cable.

It should be noted that the presence of a SCART socket on an S-VHS recorder need not mean that RGB signals are actually available: on some models, the pins reserved for the R, G and B signals (15, 11 and 7) are not connected. In the case of TV sets with a SCART socket, the RGB inputs are, fortunately, nearly always connected. None the less, this should be checked by consulting the user manual or the supplier.

The external RGB inputs of the TV set are automatically enabled when the SV7000 is connected to the SCART input.

The SV7000 is powered from a 12 VDC mains adaptor capable of supplying 300 mA. Provision has been made to protect the circuits in the SV7000 from being damaged when the external supply voltage is accidentally reversed.

Front panel controls

So far, the video industry has not been able to define a common standard for
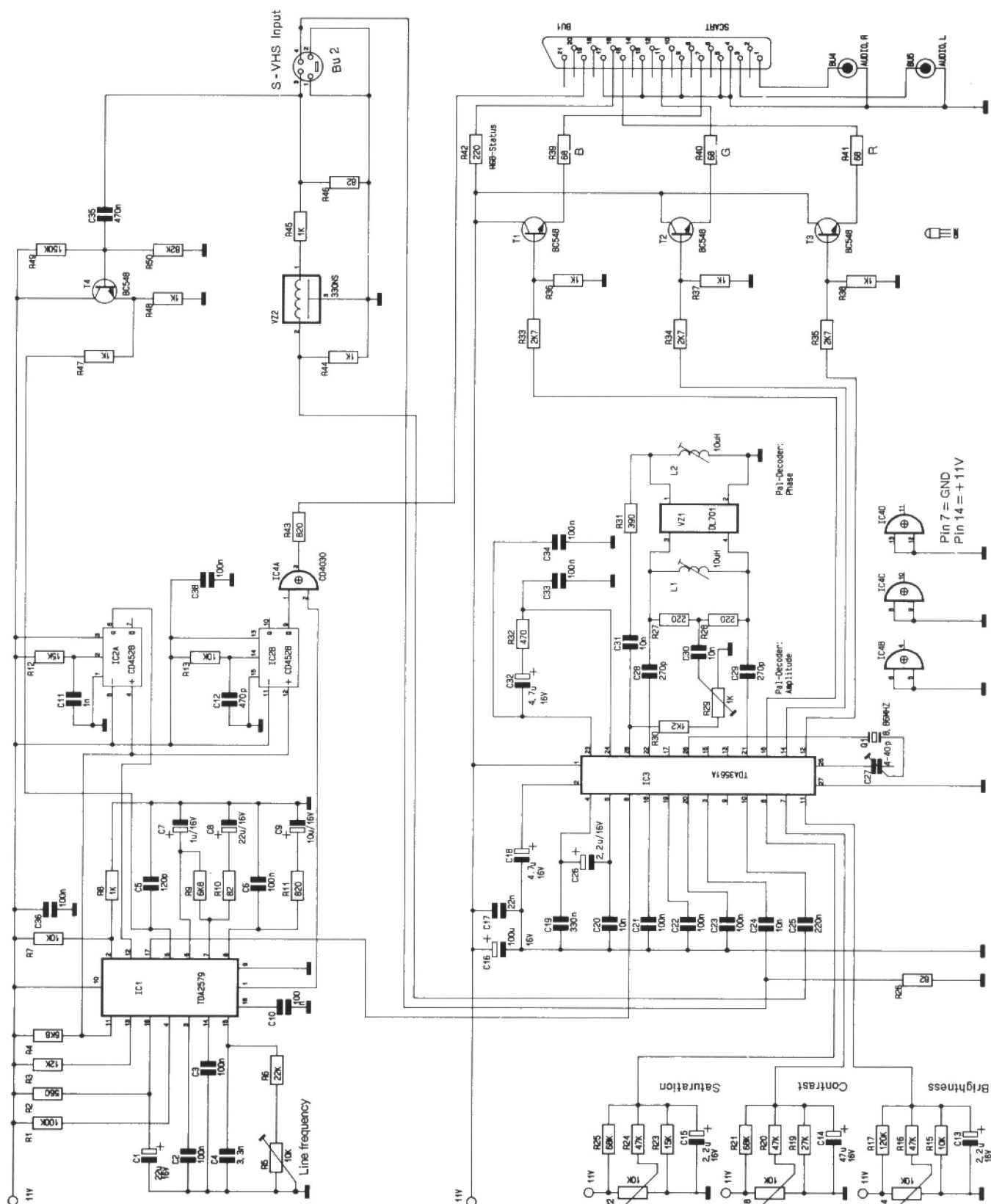
Fig. 1. Circuit diagram of the Super-VHS-to-RGB converter. A synchronization separator and a PAL decoder convert the Y-C signals from the VCR into analogue RGB and composite sync for driving a SCART input.
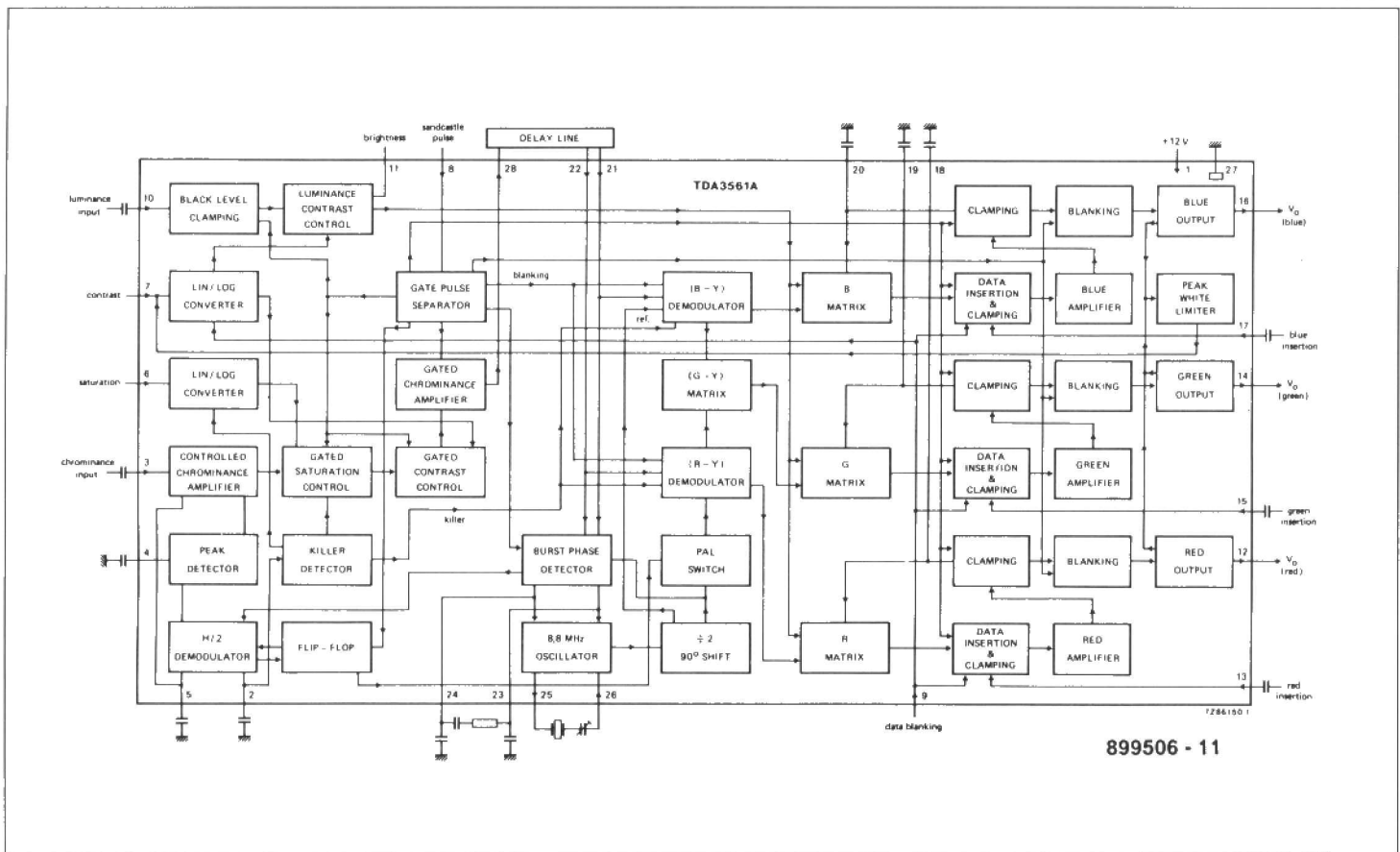
Fig. 2. Block schematic of the TDA3561A PAL decoder (*courtesy Philips Components*).

signal amplitudes. To ensure that the SV7000 gives optimum picture quality with any S-VHS recorder, it is, therefore, provided with three front panel controls to compensate level differences.

The 'Contrast' control at the left of the front panel allows the SV7000 to handle a wide range of luminance signal levels. The centre control, 'Saturation', does the same for the chrominance signal. The function of the third control, 'Brightness', speaks for itself. Together, the three controls make the SV7000 a remarkably flexible converter because it can be used with virtually any input signal level, while giving the user the opportunity of setting the picture controls to his personal taste.

Circuit description

The circuit diagram of the SV7000 is shown in Fig 1. The luminance and chrominance signals supplied by the S-VHS VCR are fed to the SV7000 via pins 4 and 3 respectively of mini-DIN socket BU2. Both signals are terminated in 82 Ω resistors, R46 and R26, to ensure correct matching to the cable and the drivers in the VCR. The chrominance and luminance signals are coupled capacitively to the corresponding inputs of single-chip PAL decoder IC3 by C24 and C25 respectively. The PAL encoder is a Type TDA3561A from Philips Components. The block schematic of this interesting chip is given in Fig. 2. The application

circuit used here is different from that in many modern TV sets to obtain a greater luminance bandwidth. In a laboratory test with a high-resolution colour monitor, the converter was found capable of correctly processing luminance signals of up to 6 MHz.

The clock oscillator on board the TDA3561A is set to work at 8.86 MHz (2 times the PAL chrominance subcarrier frequency) under the control of a quartz crystal, Q1, which is trimmed by C27. All operations internal to the chip are timed by signals derived from the 8.86 MHz clock.

PAL delay line VZ1 (a Type DL701) introduces a delay of 1 picture line (64 µs) and enables colour difference signals B−Y and R−Y to be extracted

from the chrominance signal. Phase correction is provided by inductor L2; amplitude correction by L1 and preset R29. Potentiometers R22, R18 and R14 allow colour saturation, contrast and brightness to be set to individual taste.

The decoded RGB signals available at pins 12, 14 and 16 of the TDA3561A are fed to drivers T1, T2 and T3 via series resistors R35, R34 and R33 respectively. The 68 Ω emitter resistors ensure correct matching to the respective RGB inputs on the colour TV set.

Pin 16 of the SCART connector is made high via R42 to force the TV set to connect the inputs of its RGB amplifiers to the corresponding inputs on the SCART socket.
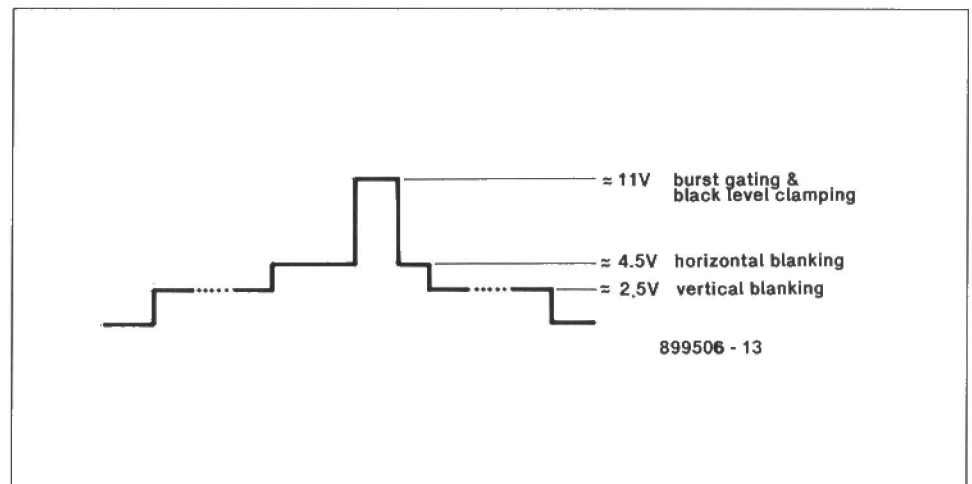
Fig. 3. The sandcastle waveform has timing and level-definition functions.

The synchronization pulses are extracted from the S-VHS luminance signal with the aid of integrated synchronization separator Type TDA2579 (IC₁), an IC typically used in conjunction with the TDA3561. Driven by buffer T₄, the TDA2579 generates separate horizontal and vertical synchronization pulses and a so-called *sandcastle* waveform (Fig. 3) which the TDA3561A needs for its internal timing, level definition and keying out of the colour burst. The sandcastle waveform available at pin 17 of the TDA2579 is fed direct to pin 8 of the TDA3561A. Exclusive-OR (XOR) gate IC₄ₐ combines the vertical synchronization pulses from pin 1 of the TDA2579 with the horizontal sync pulses from one-shots IC₂ₐ-IC₂ᵦ. The resulting composite-sync signal is fed to pin 19 of the SCART connector via series resistor R₄₃.

The 11 V regulated power supply for the S-VHS-to-RGB converter is a low-drop design around discrete components — see Fig. 4. Diode D₁ protects the converter circuit and the supply itself by causing fuse SI₁ to blow when the unregulated input voltage from the 12-15 VDC mains adaptor is accidentally reversed.

Construction

Populating the high-quality printed-circuit board (Fig. 5) and the further construction of the S-VHS-to-RGB converter is a straightforward job that will take three hours or so. Wiring is kept to a minimum by PCB-mount sockets and potentiometers.
The construction of the PCB is best started with fitting the 15 wire links. Then follow the low-profile parts, the vertically mounted parts, and the PCB-mount connectors. Before mounting the potentiometers, their spindles are cut to a length of 10 mm.

Carefully check the completed PCB before securing it to the rear panel of the enclosure with two M3×10 screws. These are inserted through the holes in the rear panel and those in the flanges of the SCART socket, and then locked with nuts at the inside of the enclosure. Remove the locking nut from the PCB-mount 3.5 mm jack receptacle for the supply voltage before inserting this in the relevant hole provided in the rear panel. Then turn the nut on the threaded shaft and secure the receptacle firmly. The front panel is fixed to the PCB in a similar fashion. Remove the nuts from the potentiometers, insert the spindles in the holes provided, and then use the nuts to secure the front panel. Lastly, fit the three knobs on to the spindles.

Now first align the circuit as detailed below.

After alignment, the converter is housed in the attractively styled, lightweight ABS enclosure supplied with the kit. Place the PCB with front and rear panel attached into the grooves provided in the

Parts list

Resistors:
R39;R40;R41=68R
R10;R26;R46=82R
R27;R28;R42=220R
R54;R55;R56=270R
R31=390R
R32=470R
R2=560R
R11;R43=820R
R8;R36;R37;R38R44;R45;R47;R48;R51;R52;
  R53=1K0
R30=1K2
R33;R34;R35=2K7
R4;R9=6K8
R7;R13;R15=10K
R3=12K
R12;R23=15K
R6;R57=22K
R19;R58=27K
R16;R20;R24=47K
R21;R25=68K
R50=82K
R1;R60=100K
R17=120K
R49=150K
R29=1K0 preset H
R5=10K preset H
R59=100K preset H
R14;R18;R22=10K potentiometer

Capacitors:
C5=120p
C28;C29=270p
C12=470p
C11=1n0
C4=3n3
C20;C24;C30;C31=10n
C17=22n
C2;C3;C6;C10;C21;C22;C23;C33;C34;C36;
  C38=100n
C25=220n
C19=330n
C35=470n
C7=1µ0; 16 V
C13;C15;C26=2µ2; 16 V
C18;C32=4µ7; 16 V
C9;C41=10µ; 16 V
C1;C8=22µ; 16 V
C14=47µ; 16 V
C16;C40=100µ; 16 V
C39=2200µ; 16 V
C27=40p trimmer
C37= not fitted

Semiconductors:
D1=1N4001
D2=5V6; 400 mW zener diode
T1...T6 incl.=BC548
T7=BD240
IC1=TDA2579
IC2=4528
IC3=TDA3561A
IC4=4030

Miscellaneous:
Q1= quartz crystal 8.86 MHz.
VZ1= DL701 PAL delay line.
VZ2= 330ns delay line.
L1;L2= 10µH adjustable inductor.
BU1= female SCART connector.
BU2= mini-DIN S-VHS socket.
BU3= 3.5 mm jack socket.
BU4;BU5= phono socket.
SI1= fuse; 0.63 A slow; with PCB-mount holder.

Fig. 4. Circuit diagram of the on-board low-drop power supply. The unregulated input voltage is obtained from a mains adaptor with 12 to 15 VDC output at about 300 mA.
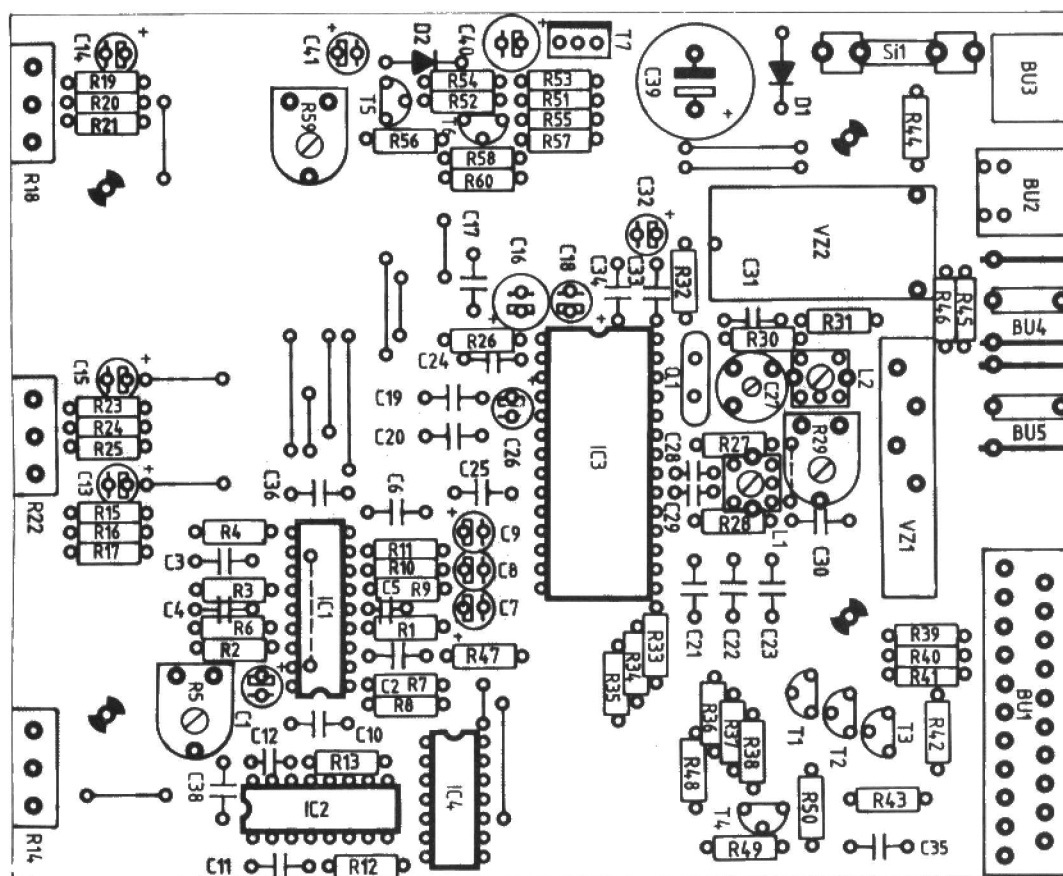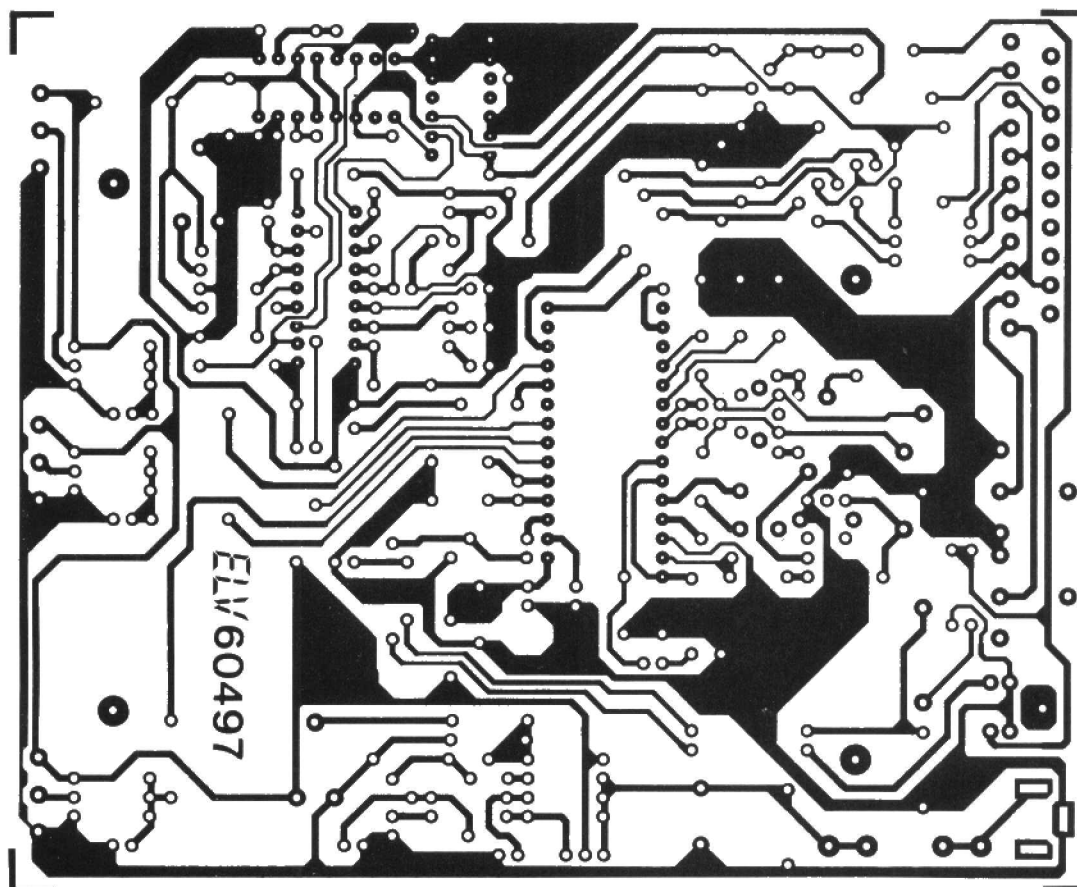
Fig. 5. Track lay-out and component mounting plan.
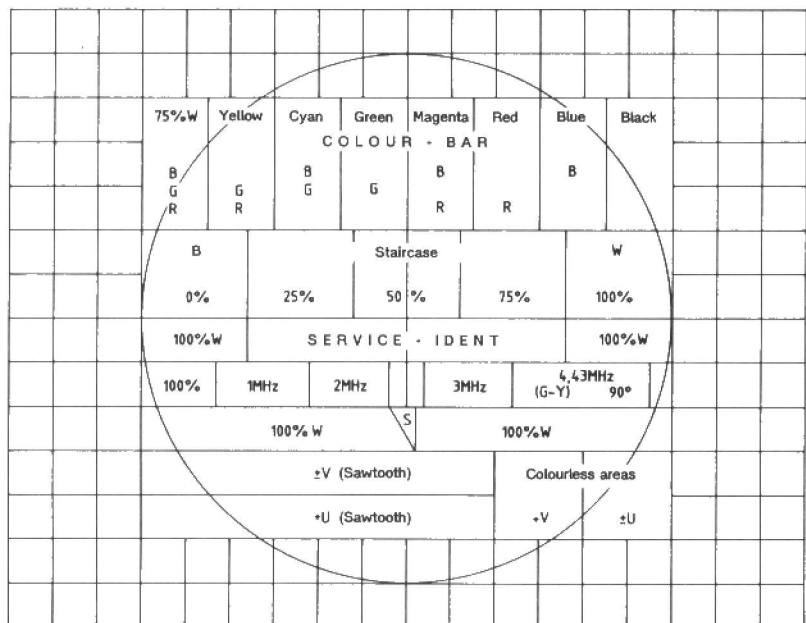
Fig. 6. Functional analysis of the test signals available in the Type FuBK electronic test chart.

bottom section of the enclosure. Then mount the top section, and secure it to the lower section with the aid of two long screws inserted from the underside.

Setting up

The S-VHS-to-RGB converter is relatively simple to align with the aid of an S-VHS recorder, a recording of an electronic test chart, and a TV set or monitor with SCART input.

Start with a separate test of the power supply and adjust preset R59 for an output voltage of 11 V.

Make a recording of half an hour or so of an electronic test chart. The FuBK (Fig. 6) and PM5534 (Fig. 7) test charts are given here for reference.

Connect the SV7000 to the recorder, the TV set and the mains adaptor. Set the trimmer, the two presets and the three picture controls to the centre of their travel. Start the tape. If necessary, synchronize the picture by adjusting R5. The colour may fail at this stage, but can be restored by carefully adjusting trimmer C27.

The PAL decoder is aligned by carefully
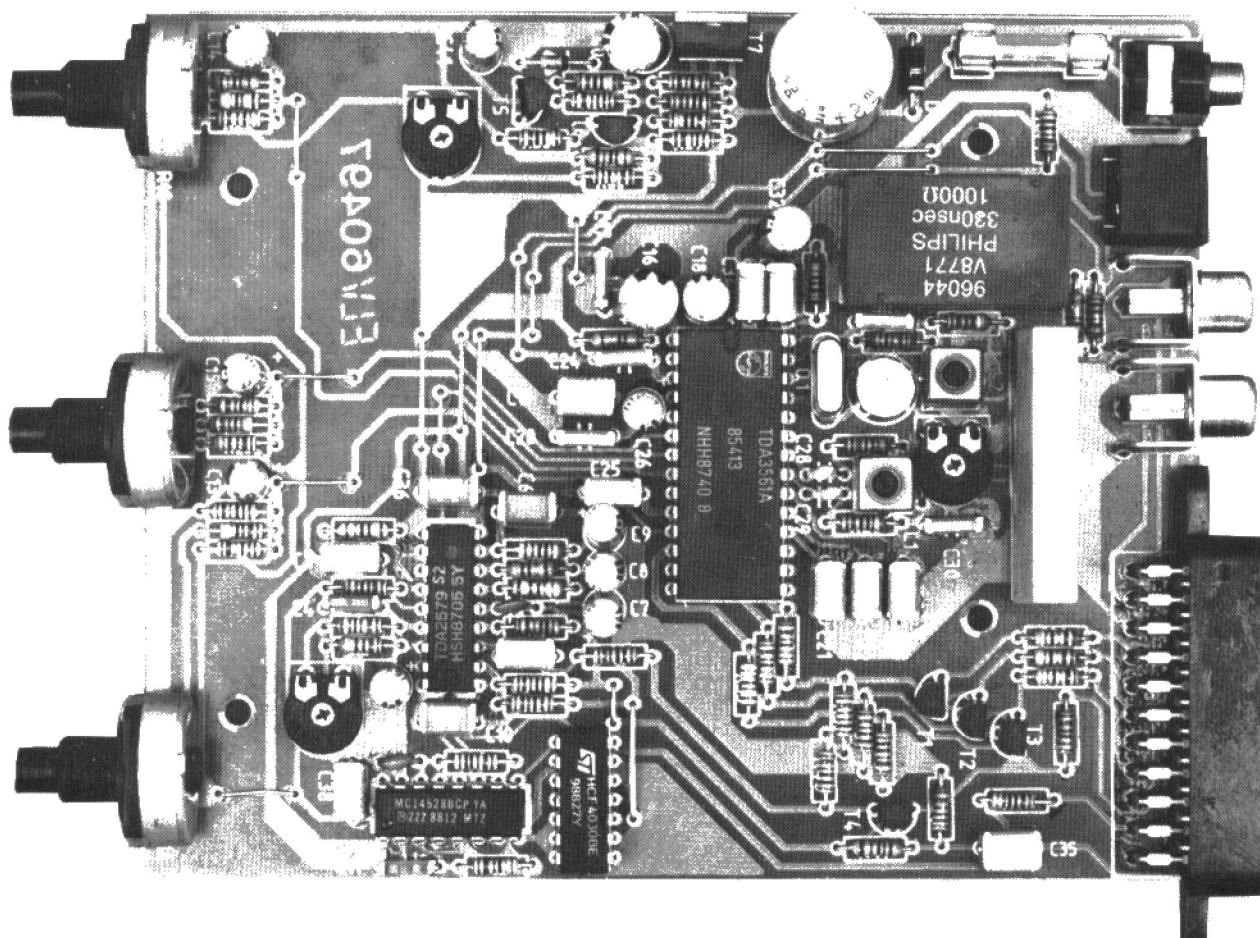
Top view of the completed printed-circuit board. Note the use of PCB-mount potentiometers and sockets.

looking at the picture areas reserved for the sawtooth ±V and ±C (FuBK/PM5534: ±U) test signals and the colourless areas ±V and ±C (FuBK/PM5534: ±U). Any coloured horizontal lines or other patterning effects noticed in these areas point to amplitude errors that may be corrected by adjusting R_{29} and L_1. Similarly, when horizontal bars are noted in the G–Y area, the causative phase errors may be corrected by adjusting L_2. The picture quality is then optimized by small, alternate, corrective alignments of the amplitude (R_{29}-L_1) and phase (L_2). The inductors must be aligned with an insulated trimming tool. Never use a metal screwdriver because this changes the inductance of the coil while the core is being adjusted.

This concludes the setting up of the SV7000 S-VHS-to-RGB decoder.

Final notes

The RGB-status is forced in a fairly crude way by resistor R_{42}, and causes the TV set to switch to RGB input irrespective of whether a programme is played back on the VCR or not. This may be remedied by installing the simple extension circuit of Fig. 9, which does not pull pin 16 of the SCART connector high until synchronization pulses are detected.

The PLL circuit internal to the synchronization separator may not be fast enough to follow tape speed variations that occur in certain types of camcorders. Fortunately, this problem is solved simply by the fitting of a 180 kΩ resistor in parallel with C_{10}.

The vertical synchronization signal does not contain the so-called *back-porch equalization pulses*. On a few types of TV set or monitor, the absence of these pulses may cause slight instability or deforming effects near the top of the picture.

Note: we regret that photocopies and/or films of PCB Type ELV60497 can not be supplied through the Readers' Services.

A complete kit of parts for the SVR7000 S-VHS-to-RGB converter, which is designed in West-Germany, is available from the designers' exclusive worldwide distributors (regrettably not in the USA and Canada):

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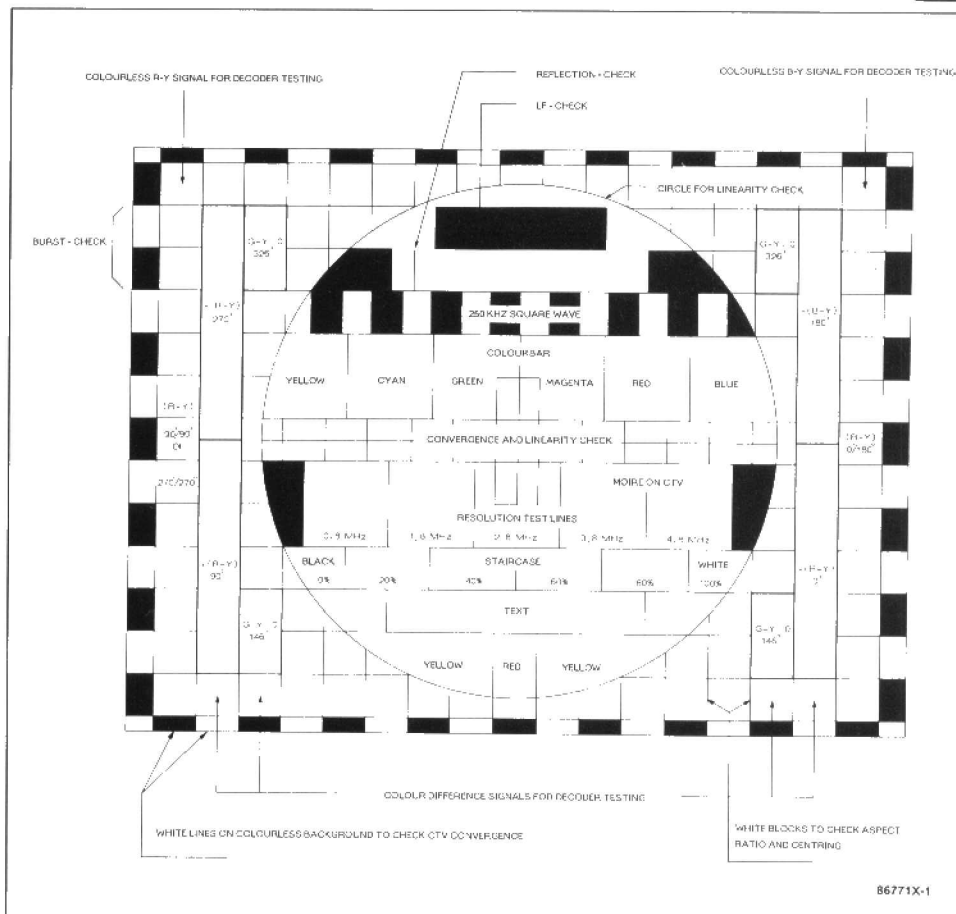
Fig. 8. Functional analysis of the test signals available in the Type PM5534 electronic test chart.
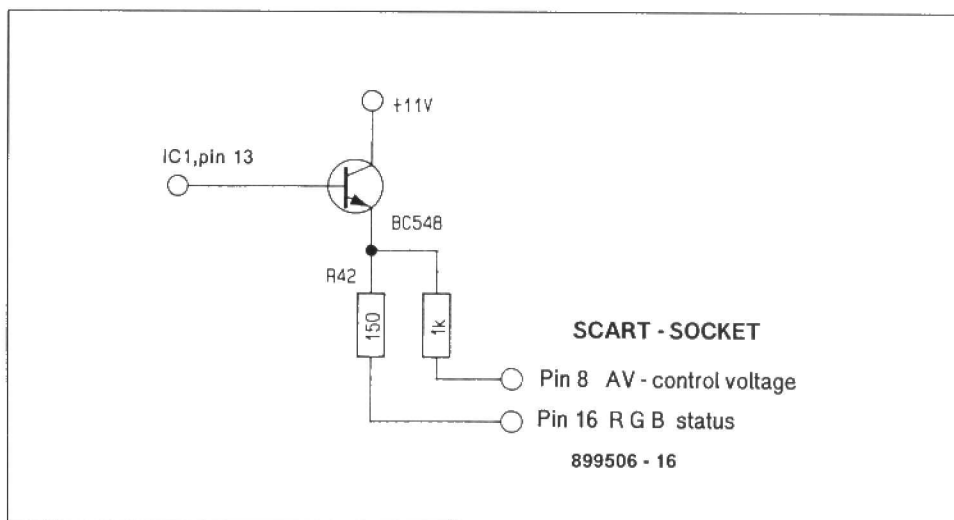
Fig. 9. Modified RGB status driver that switches the TV set to RGB (=S-VHS VCR) input when synchronization pulses are detected.
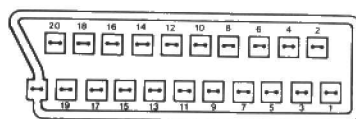
SCART connectors
Virtually all modern video equipment is fitted with a SCART connector. The pin assigment is uniform among manufacturers, but in some cases not all pins are used.

pin	signal		
1	audio out R or Ch. 2	12	not used
2	audio in R or Ch. 2	13	ground (red)
3	audio out L or Ch. 1 (or mono)	14	not used
4	ground (audio)	15	red
5	ground (blue)	16	blanking (active high)
6	audio in L or Ch. 1 (or mono)	17	ground (video)
7	blue	18	ground (blanking)
8	control voltage	19	CVBS out
9	ground (green)	20	CVBS in
10	not used	21	conector shield and/or ground
11	green		

Fig. 10. For reference: pinning of the SCART socket.

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